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USSR Report

ELECTRONICS AND ELECTRICAL ENGINEERING

(FOUO 17/80)

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18 December 1980

USSR REPORT

ELECTRONICS AND ELECTRICAL ENGINEERING

(FOUO 17/80)

CONTENTS

 [III - USSR - 21E S&T FOUO]

ANTENNAS

UDC 621.391.2:621.396.667.49

ESTIMATING THE INFLUENCE OF BANDWIDTH ON THE POSSIBILITY OF ADAPTIVE SIGNAL PROCESSING IN ANTENNA ARRAYS

Moscow RADIOTEKHNIKA I ELEKTRONIKA in Russian Vol 25, No 7, Jul 80 pp 1531-1533 manuscript received 10 Apr 79

[Paper by M.V. Ratynskiy]

[Text] 1. Adaptive processing of narrow band signals in an m-element antenna array (a weighted summing configuration with adaptive control of the weights) makes it possible in principle to completely suppress the signals from m - 1 external sources [1].

For the optimum solution which minimizes the average power at the output of a phased antenna array (PAR), this assertion follows directly from the fact that the correlation matrix of the input signals of the PAR, without taking into account the internal noise of the array elements, remains degenerate, while the number of external signal sources q remains less than the number of variable weighting coefficients m. In this case, the minimum of the average output power is governed by the minimal (zero value for a degenerate matrix) eigenvalue of the correlation matrix of the input signal, while any linear combination of eigenvectors of the correlation matrix which correspond to its zero eigenvalue can serve as the optimal weighting vector.

The indicated assertion, however, is strictly justified only for signals with spectra in the form of δ-functions. It can be shown, and this will be done below, that with a finite spectral width, the correlation matrix of the input signal proves to be fundamentally nondegenerate (although sometimes extremely poorly conditioned), even when $q < m$, starting at $q = 1$. For this reason, the depth of signal suppression where the signals have a finite spectral width always proves to be limited.

The problem of quantitatively estimating the possible depth of signal suppression of signals with a finite bandwidth in adaptive PAR's of the type considered here is treated in this paper. The major steps in the solution is accounting for the waveform and bandwidth of the signals being processed when constructing the correlation matrix. When such a matrix has been constructed, the minimum of the average power at the output of the PAR, P_m, is quantitatively determined by its minimal eigenvalue, which can be found by well known numerical methods. The maximum possible suppressions depth is then defined as the ratio of P_m tc the original power:

(1) $$P = W^T B W^*,$$

where W is the original weighting vector with a unit euclidian valuation; T is the transposition sign; * is the complex conjugate sign; B is the correlation matrix of the PAR input signals.

2. We shall start with the correlation matrix:

(2) $$B(\tau) = \frac{1}{2}\overline{X(t)X^H(t-\tau)},$$

where X(t) is an m-dimensional vector of the input signals (complex envelopes: KO) of the PAR; H is the Hermitian conjugacy sign (transposition and complex conjugacy); the bar is the sign for averaging over the set. The elements of the matrix $B(\tau)$ are the autocorrelation functions of the complex envelopes (the AKFKO - the elements of the main diagonal) and the cross-correlation functions of the complex envelopes (VKFKO - the remaining elements) of the input signals of the elements of the array. By subjecting equation (2) to a Fourier transform, we obtain the spectral matrix:

(3) $$A(\omega) = \frac{1}{2}\int_{-\infty}^{\infty} B(\tau) e^{-j\omega\tau}\, d\tau,$$

the elements of which are the energy spectra of the complex envelopes (ESKO's are the elements of the main diagonal) and the mutual power spectra of the complex envelopes (VESKO are the remaining elements) of the input signals of the elements of the PAR.

To simplify the analysis, we shall assume that the vector X does not include the internal noise of the array components, while the external signal is due to one source (q = 1). The second limitation below will be moved, while the first, obviously, leads to a reduction in all of the eigenvalues of the matrix B(0) by the amount of the dispersion of the internal noise of one array element (with a euclidian valuation of the weighting vector equal to unity). Then, if the input signals are narrow band ones with a center frequency of the spectrum of ω_0, it is not difficult to prove that the mutual energy spectra of the complex envelopes of the input signals of the PAR elements with numbers i and k are related to the energy spectra of the complex envelopes of the input signal of any of them by the relationship:

(4)
$$S_{ik}(\omega)=S_{ii}(\omega)\,e^{j(\omega_0+\omega)\Delta t_{ik}},$$

where $\Delta t_{ik} = \Delta t_k - \Delta t_i$ is the difference in the delays of the input signals of the i-th and k-th elements of the PAR relative to the input signal of the first element. In fact, the mutual correlation and autocorrelation functions of the complex envelopes of the input signals of the corresponding elements are related by the expression:

$$B_{ik}(\tau)=B_{ii}(\tau+\Delta t_{ik})\,e^{j\omega_0\Delta t_{ik}},$$

from which (4) follows after integrating with respect to (3). The element b_{ik} of the desired correlation matrix $B = B(0)$, included in the hermitian of the form of type (1), can now be obtained by means of a back Fourier transform of the corresponding element of spectral matrix (3), taking (4) into account:

(5)
$$b_{ik}=\frac{1}{\pi}\int_{-\infty}^{\infty}S_{ik}(\omega)\,d\omega.$$

We shall now consider three cases separately, where the form of the spectrum of the input signals has the form of a δ-function, a rectangle and a gaussian curve. In the first case:

$$S_{ii}(\omega)=\sigma^2\delta(\omega)$$

and

(6)
$$b_{ik_\delta}=\frac{\sigma^2}{\pi}\,e^{j\omega_0\Delta t_{ik}}.$$

In the second case:

$$S_{ii}(\omega)=\begin{cases}\sigma^2/\Delta\omega_n, & |\omega|<\Delta\omega_n/2,\\ 0, & |\omega|>\Delta\omega_n/2\end{cases}$$

(7)
$$b_{ik_n}=\frac{\sin(\Delta\omega_n\Delta t_{ik}/2)}{\Delta\omega_n\Delta t_{ik}/2}\,b_{ik_\delta}.$$

In the third case:

$$S_{ii}(\omega) = (\sigma^2/\Delta\omega_\Gamma)\, e^{-\pi\omega^2/\Delta\omega_\Gamma^2}$$

$$b_{ik_\Gamma} = e^{-\Delta\omega_\Gamma^2 \Delta t_{ik}^2/4\pi}\, b_{ik_\delta}.$$

(8)

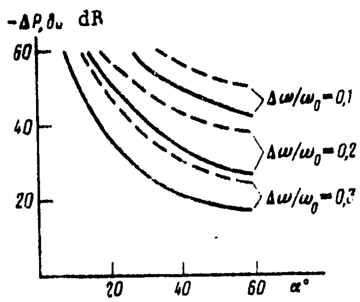

Figure 1.

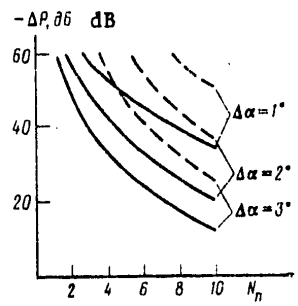

Рис. 2 ($\alpha_{нач}$=20°, $\Delta\omega/\omega_0$=0,1)

Figure 2.

It follows from the results obtained that for a finite spectral width, the linear function of the rows (or columns) of the correlation matrix is violated, i.e., the latter becomes nondegenerate [2], something which also indicates the fundamental impossibility of complete suppression of signals with a finite spectrum width. Calculations show that quantitative estimates for signals with rectangular and gaussian spectral forms prove to be rather close. For this reason, one can speak in practical terms about estimates for signals with a bandwidth of $\Delta\omega = \Delta\omega_\Pi = \Delta\omega_\Gamma$, without specifying the form of the spectrum, but assuming that it is rather simple (a single peaked curve).

Returning to the case of several independent external signal sources, it is not difficult to convince oneself that the resulting correlation matrix is equal to the sum of the component matrices (depending on the number of sources), each of which is constructed for its own source taking into account the width of its spectrum, i.e., in accordance with (6)-(8).

3. We shall give examples of estimates obtained by the method considered here.

The curves for the maximum possible suppression depth ΔP as a function of the direction α to the signal source, read out from a normal to the aperture, for a linear equally spaced array with a uniform amplitude distribution for a mounting step of the radiators equal to the wavelength, are shown in Figure 1 for the case where the number of array elements is m = 4 (the solid curves) and m = 6 (the dashed lines), for three values of the bandwidth $\Delta\omega/\omega_0 = 0.1$, 0.2 and 0.3. The functions shown here confirm that the possibilities for suppression prove to be more limited in step with an increase in the bandwidth $\Delta\omega$. At the same time, the possibilities for suppression somewhat increase when the number of array elements increases.

Figure 2 illustrates the possibility for the suppression of signals from several sources, positioned at angular distances $\Delta\alpha$ from each other. For the first source, $\alpha = 20°$, and the array is a linear equally spaced one; the solid lines correspond to m = 4 and the dashed lines to m = 6. It can be seen from the graphs that the suppression possibilities fall off markedly with an increase in the number of sources, N_π, and depend substantially on their mutual angular positioning.

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[318-8225]

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CSO: 1860

CERTAIN ASPECTS OF PHOTOGRAPHY, MOTION PICTURES AND TELEVISION

UDC 53.087.9

TELEVISION OPTICAL VELOCITY DETECTOR

Leningrad IVUZ, PRIBOROSTROYENIYE in Russian Vol 23, No 7, 1980 pp 76-78 manuscript received 9 Jul 79

[Article by N.F. Gusarov and S.A. Sukhoparov, Leningrad Institute of Precision Mechanics and Optics]

[Text] The results are given of an experimental investigation of a television optical detector, developed by the authors, of the rates of motion of a random field of brightness.

Monitoring the rates of motion of a random field of brightness (e.g., of hot rolled metal) remains a topical problem up to the present time [1, 2]. One of the variants of the "light-signal" converter used in systems for this kind of monitoring can be an instantaneous-action television tube--a dissector tube, the advantages of which are the relative simplicity of its circuitry, high reliability, and independence of its noise characteristics on the resolution standard employed [3].

The authors have developed and tested a detector of the rate of motion of a random brightness field based on a series-produced LI-605-1 dissector tube. The signal in the dissector tube's load represents the result of conversion of the space spectrum of the distribution of illumination intensity in the photocathode into the time region according to the scanning law. Since the space spectrum of the monitored field, and, consequently, its time transformation are not known beforehand, then the differential method is used for the purpose of isolating information on the rate of motion of the field. Here scanning which is symmetric for the period is used (e.g., triangular) [4], and processing of the signal takes place in relation to the two channels being compared, corresponding to the forward and reverse scanning operations.

A structural diagram of the detector developed is presented in fig 1. The input optics (O) convert the angular distribution of energy brightness of the field being monitored into the structure and intensity of the image on the photocathode of the dissector tube (D). The signal from the dissector tube's load is fed to a shaper amplifier (UF), where it is converted into a train of pulses whose time position corresponds to the position of black and white differentials in the image in relation to the scanning reference point. A commutator (K) divides the pulse train into two channels corresponding to the forward and reverse scanning operations. From the sweep and synchronizing generator (GRS) the sweep voltage is supplied to the deflection coils of the focusing and deflection system and synchronization of the entire system is carried out. The difference in the number of pulses counted during the forward and reverse scanning operations, Δn, is computed in the reversible counter (RS) and serves as a measure of the rate of movement of the monitored field, and the sign of this difference corresponds to the direction of motion.

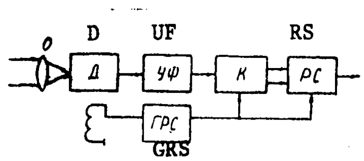

Figure 1. Structural Diagram of Detector

Although the detector is designed for monitoring the rate of motion of random fields of brightness, it is feasible to estimate its conversion transconductance by means of the simplest--the regular--model of a field of brightness. Let there be projected onto the photocathode (FK) an image of a line-type optical focusing pattern, whose transmission (in coordinates relating to the photocathode) equals $E = E_0 \sin \Omega x$. The image moves at a constant rate of $\vec{v}$ over the photocathode along the x axis; scanning also takes place along the x axis. Let us assume for simplicity that the scanning rate in the forward and reverse operations is steady and equals $\pm v_r$ (which corresponds to triangular scanning). Then the voltage separated in the dissector tube's load during each half-cycle equals

$$U_1 = \varepsilon E_0 \sin[\Omega(v_r - v)t] \quad \text{при} \quad \vec{v}_r \uparrow\uparrow \vec{v};$$
$$U_2 = \varepsilon E_0 \sin[\Omega(v_r + v)t] \quad \text{при} \quad \vec{v}_r \uparrow\downarrow \vec{v}, \qquad (1)$$

where ε is the sensitivity of the photocathode, Ω is the spatial density of the image and t is current time.

The repetition rate of pulses formed in the shaper amplifier equals

$$T_1 = \frac{2\pi}{\Omega(v_r - v)}; \quad T_2 = \frac{2\pi}{\Omega(v_r + v)}, \qquad (2)$$

and the number of these pulses during a scanning half-cycle, $T/2 = \pi/\nu$, equals:

$$n_1 = \frac{T/2}{T_1}; \quad n_2 = \frac{T/2}{T_2}. \tag{3}$$

Substituting (2) in (3) and taking into account the fact that the scanning cycle can be expressed in terms of the scanning rate, v_r , and the length of a line of resolution, d , $T = 2d/v_r$, and the linear rate of motion of the image can be determined by the angular velocity of the monitored field, ω , and by the focal length of the lens, f ,

$$v = \omega f, \tag{4}$$

we get an expression for Δn :

$$\Delta n = \frac{\Omega f}{\nu} \omega. \tag{5}$$

If the field has a random structure, then Ω represents the mean spatial density of the image presented to the photocathode. From (5) it is possible to draw a number of conclusions:

1. The detector described is not an absolute gauge of velocity, but only an indicator of its relative value and sign.

2. The sensitivity and conversion transconductance are best when working with fine-structure objects with high values of Ω .

3. It is possible to vary the range of monitored velocities over a wide range by selection of the scanning rate, ν .

In fig 2 is given the experimental dependence $\Delta n\nu = \phi(\omega)$. Tests were performed for two types of surfaces: 1--a line-type optical focusing pattern ($\Omega = 10$ cm^{-1} and contrast $K_0 = 0.5$); 2--a random "mottled" surface ($\Omega = 6$ cm^{-1} and $K_0 = 0.5$). The dissector tube operated in the single-line mode with a scanning rate of $\nu = 2.5$ Hz and the direction of the scanning line agreed with the direction of the motion monitored, the signal-to-noise ratio in the dissector tube's output equaled $\Psi = 47$ and the focal length of the lens equaled $f = 480$ mm . From the graphs it is obvious that in both cases the dependences arrived at are approximated sufficiently accurately by a linear law. Consequently, for the purpose of arriving at the real velocity it is sufficient to multiply the measured value by some factor constant for the class of fields in question.

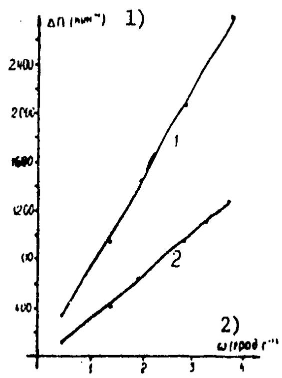

Figure 2. Results of Experimental Investigation of Detector: 1--for regular structure; 2--for random structure

Key:
1. min^{-1}
2. $deg \cdot s^{-1}$

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COMPONENTS AND CIRCUIT ELEMENTS, WAVEGUIDES, CAVITY RESONATORS AND FILTERS

UDC 62-501.12-501.7

METHOD OF SYNTHESIZING ADAPTIVE SUBOPTIMAL FILTERS

Leningrad IVUZ, PRIBOROSTROYENIYE in Russian Vol 23, No 7, 1980 pp 17-21
manuscript received 18 Jun 79

[Article by G.F. Filaretov and V.S. Kumenko, Moscow Power Engineering Institute]

[Text] A method is suggested for synthesizing adaptive suboptimal filters for an estimate of the state vector of a dynamic system based on a linear Kalman filter. In the method are utilized the principles of designing self-tuned systems with a model. An example is given of the synthesis of a second-order filter along with the results of simulation.

In recent times special attention has been attracted by multidimensional Kalman filters [1-3] for an estimate of the state of a dynamic system. They have a recurrence form convenient for implementation on a digital computer and theoretically make possible the determination of estimates with minimal variance between all linear unbiased estimates. However, in processing measurements with special-purpose digital computers in real time the employment of optimal Kalman filters encounters a number of difficulties.

First, these include register errors which result in deviation of the machine solution from the mathematical and even in the divergence of filters [5]. Second, multidimensional filtering requires considerable expenditures of machine time [4, 8]; as investigations have demonstrated, the time for processing measurements is proportional to the cube of the dimensionality of the state vector of the dynamic system. Third, an analysis of the sensitivity of Kalman filters [1, 6, 7] has demonstrated that disparity between the suggested equation of motion according to which the filter is designed and the actual change in the system's state vector results in rapid divergence of the filter.

On the basis of the above, the aspiration of investigators to find some compromise solution in selecting a filter for measurement processing is understandable. Many investigators [1, 4, 5, 8-10] have attempted, on account of the slight losses in optimality in specific problems, to employ the Kalman filtering procedure for designing simpler stable suboptimal filters.

In this paper is considered the class of problems in which the real motion of the dynamic system does not agree with the equations of motion according to which the Kalman filter is designed. Such a problem arises if all "forces" responsible for evolution of the dynamic system are unknown, or in complicated problems when for the purpose of the simplification of computations one is restricted to an approximate description of the system.

For this class of problems it is suggested that adaptive suboptimal filters be used, the design principles of which are given in [11, 12].

Let the motion of the dynamic system be described by the equation

$$x(i+1)=A_i\cdot x(i)+\Gamma_i w(i),$$
$$E\{w(i)\}=0,\ \ E\{w(i)\cdot w^T(j)\}=Q_i\delta_{ij}, \tag{1}$$

and let there be a measurement vector

$$y(i)=H_i x(i)+\xi(i),$$
$$E\{\xi(i)\}=0,\ \ E\{\xi(i)\cdot\xi^T(j)\}=R_i\delta_{ij},\ \ E\{w(i)\cdot\xi^T(j)\}=0, \tag{2}$$

where $E\{\cdot\}$ is the mathematical expectation operation; H_i, A_i and Γ_i are the matrices of transformations; and Q_i and R_i are the matrices of covariants of random vector quantities $\omega(i)$ and $\xi(i)$, respectively.

In this case a Kalman filter gives an unbiased estimate of the state vector, $x(i)$, with minimal variance. This estimate is determined by the equations

$$\hat{x}(i+1)=A_i\hat{x}(i)+K_{i+1}\cdot[y(i+1)-H_{i+1}A_i\hat{x}(i)], \tag{3}$$

$$K_i=P_iH_i^TR_i^{-1} \tag{4}$$

$$P_i=[M_i^{-1}+H_i^TR_i^{-1}H_i]^{-1} \tag{5}$$

$$M_{i+1}=A_iP_iA_i^T+\Gamma_iQ_i\Gamma_i^T, \tag{6}$$

where K_i is the matrix of weighting factors,

$$P_i = E\{[x(i) - \hat{x}(i)][x(i) - \hat{x}(i)]^T\},$$
$$M_i = E\{[x(i) - A_{i-1}\hat{x}(i-1)][x(i) - A_{i-1}\hat{x}(i-1)]^T\},$$

(7)

P_i is the matrix of covariants of the error in the estimate and M_i is the matrix of covariants of the extrapolation error. The initial estimate, x(0) , and matrix, P_0 , are assumed to be assigned.

The method of synthesizing an adaptive filter consists in finding the systematic error between the real motion and the hypothesis for designing the Kalman filter and in including the value of this error in equations (5) to (7) for the purpose of calculating the matrix of weighting factors, K_i . In other words, it is necessary to select the state vector of the new dynamic system and from measurements

$$\varepsilon(i) = y(i) - H_i A_{i-1} \cdot \hat{x}(i-1) \tag{8}$$

to construct a secondary filter for estimating the systematic error.

It is suggested that as the state vector in designing the secondary filter be used either the vector of the estimate error

$$e(i) = x(i) - \hat{x}(i), \tag{9}$$

or the vector of the extrapolation error

$$u(i) = x(i) - A_{i-1}\hat{x}(i-1). \tag{10}$$

Let us consider the second approach to synthesizing an adaptive filter.

First it is necessary to arrive at an equation describing the evolution of state vector u(i) and to find an equation for the relationship between the measurement vector, $\varepsilon(i)$, and the state vector. From equation (10), utilizing (1) to (3), we have

$$u(i+1) = A_i[J - K_i H_i]u(i) + \Gamma_i w(i) - A_i K_i \xi(i). \tag{11}$$

From equation (8) and (2) we get

$$\varepsilon(i) = H_i u(i) + \xi(i). \tag{12}$$

To equations (11) and (12) it is possible to apply the well-known [3] Kalman filtering procedure for the correlated noise in the equations.

Let us write the estimate and extrapolation equations in the following form:

$$\hat{u}(i)=\bar{u}(i)+K_i^0[\varepsilon(i)-H_i^0\bar{u}(i)], \tag{13}$$

$$\hat{u}(i+1)=A_i^0\hat{u}(i)+D_i[\varepsilon(i)-H_i^0\hat{u}(i)]. \tag{14}$$

Weighting matrices K_i^0 and D_i can be determined by the recurrence equations:

$$D_i=\Gamma_i^0 S_i (R_i^0)^{-1} \tag{15}$$

$$\Pi_i=[(M_i^0)^{-1}+(H_i^0)^T(R_i^0)^{-1}H_i^0]^{-1} \tag{16}$$

$$K_i^0=\Pi_i(H_i^0)^T(R_i^0)^{-1}, \tag{17}$$

$$M_{i+1}^0=(A_i^0-D_iH_i^0)\Pi_i(A_i^0-D_iH_i^0)^T+\Gamma_i^0Q_i^0(\Gamma^0)^T-D_iR_i^0D_i^T, \tag{18}$$

where

$$A_i^0=A_i[J-K_iH_i],\ H_i^0=H_i, \tag{19}$$

$$\Gamma_i^0=-A_iK_i,\ Q_i^0=R_i^0=S_i=R_i. \tag{20}$$

In addition, in deriving these equations it is possible to assume that $\Gamma_i = 0$. This assumption does not restrict the class of problems which can be considered, since the amount of the error is included in the estimate vector, $\hat{u}(i)$.

By means of equations (19) and (20) system of equations (15) to (18) is brought to the form:

$$D_i=-A_iK_i \tag{21}$$

$$\Pi_i=[(M_i^0)^{-1}+H_i^TR_i^{-1}H_i]^{-1} \tag{22}$$

$$K_i^0=\Pi_iH_i^TR_i^{-1} \tag{23}$$

$$M^0_{l+1} = A_l \Pi_l A^T_l \qquad (24)$$

Now it is necessary to introduce the amount of the estimate, $\hat{u}(i)$, into system of equations (5) to (7). Since magnitude $\hat{u}(i)$ represents the systematic extrapolation error and matrix M_i is the matrix of covariants of the extrapolation error, then equations (6) to (7) are changed in the following manner:

$$P_l = [M^{-1}_{al} + H^T_l R^{-1}_l H_l]^{-1} \qquad (25)$$

$$M_{a\,l+1} = A_l P_l A^T_l + \hat{u}(l)\,[\hat{u}(l)]^T \qquad (26)$$

where subscript a designates an adaptive matrix.

Thus, the set of recurrence equations (3), (5) and (10) to (26) defines an adaptive suboptimal filter making it possible to estimate the state vector of dynamic system x(i) with the existence of a systematic error in equation of motion (1).

Example

Let it be required to estimate coordinate x(i) and its first derivative v(i) from measurements y(i) . Here, the motion of the system can be described by a difference equation both of the first order (even straight-line motion) and of the second order (evenly accelerated motion). Since it is necessary to estimate only the coordinate and the rate of its change, we will construct an estimate algorithm from the condition of even motion. When shifting to evenly accelerated motion the parameters of the algorithm require readjustment; therefore, an adaptive algorithm is required for the estimate.

The motion and measurement equations take on the form

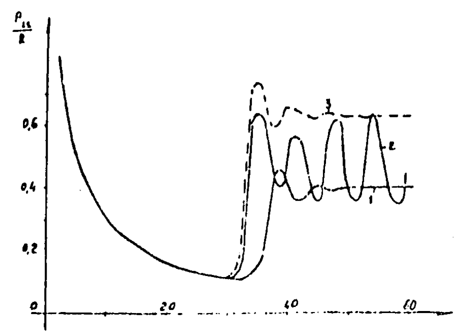
$$\begin{bmatrix} x(l+1) \\ v(l+1) \end{bmatrix} = A \begin{bmatrix} x(l) \\ v(l) \end{bmatrix}; \quad y(l) = H \begin{bmatrix} x(l) \\ v(l) \end{bmatrix} + \xi(l),$$

where

$$A = \begin{bmatrix} 1 & T \\ 0 & 1 \end{bmatrix}, \quad H = [1 \quad 0].$$

In designing an adaptive filter, instead of the equations for a secondary filter, (13) to (19), let us employ a filter with an effective finite memory [2, 5] and let us assume that the weighting factor matrix, $K°$, is constant.

First let the system move at a constant speed and then let acceleration be evidenced at a certain moment of time. In fig 1 are given the reactions of an adaptive filter with various amounts of acceleration. The characteristics of the adaptive filter for this problem are determined by ratio γ , equaling the acceleration divided by the root-mean-square measurement error. In this figure curve 1 corresponds to $\gamma = 0.05$ and 2 to $\gamma = 0.2$. It is obvious that in case 2 the filter becomes unstable with selected k_1 and k_2 ; however, if the secondary filter is made with quicker response (e.g., $k_1 = 0.7$ and $k_2 = 0.3$), then the adaptive filter maintains its stability--curve 3. If the filter is unstable with any $K°$ matrix, then this means that it is impossible to provide for filtering by means of a second-order filter. In this case it is necessary to use filters of a higher order.



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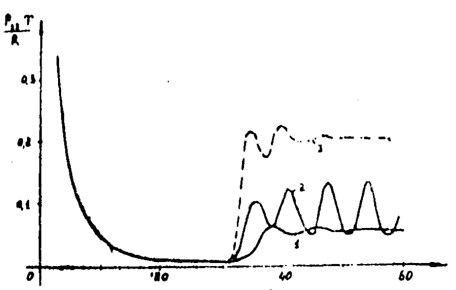

b)

Figure 1. Evolution of Elements of Matrix of Covariants, P , with a Change in the Nature of Motion: 1--$\gamma = 0.05$, $k_1 = 0.39$

and $k_2 = 0.067$; 2—$\gamma = 0.2$, $k_1 = 0.39$ and $k_2 = 0.067$;
3—$\gamma = 0.2$, $k_1 = 0.7$ and $k_2 = 0.3$

1. In this paper a method has been suggested for synthesizing adaptive suboptimal filters making it possible to estimate the state vector of a nonstationary dynamic system without revealing the very case of nonstationarity.

2. The application of these filters will make it possible to simplify algorithms for estimating the state vector, which is especially important in implementing algorithms on short-bit-configuration digital computers.

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UDC 621.391.2

ON AN ANALYSIS OF THE SENSITIVITY OF A TWO-DIMENSIONAL KALMAN FILTER IN THE PROBLEM OF STATISTICAL MONITORING OF CHANNEL QUALITY

Moscow RADIOTEKHNIKA I ELEKTRONIKA in Russian Vol 25, No 7, Jul 80 pp 1537-1540 manuscript received 21 Feb 78 after revision 3 Jan 80

[Paper by V.G. Koveshnikov and A.N. Sharov]

[Text] In a whole series of applied problems in communications theory which come up when analyzing the quality of the working channel of an information system, it is necessary to obtain a reliable estimate of the power parameters of the signal and noise present at the input to the receiver in the form of an additive mixture. An algorithm which is optimal in the sense of a mean square criterion for the separate estimation and prediction of the signal and interference levels in a working channel, which takes the form of a two-dimensional modified Kalman filter, was synthesized in paper [3] using the methods of the theory of state variables [1, 2], assuming complete apriori information on the correlation properties of the Markov signals and interference. It is natural that in those cases where the actual dynamics of the processes being studied do not correspond to the apriori data on the correlation properties and the interference, incorporated in such an analyzer-extrapolator, an increase in the measurement error is to be anticipated when monitoring channel quality.

We shall analyze the sensitivity of the analyzer-extrapolator for channel quality, which was synthesized in [3], where this analysis makes it possible to estimate the effeciency of these kinds of analyzers-extrapolators under conditions of incomplete apriori information on the correlation properties of the interference.

We shall compute the error in the measurements based on the procedure given in [4], taking into account the assumptions made in paper [5]. In this case, we obtain the following expressions for the estimation of the sensitivity of the analyzer-extrapolator synthesized in paper [3]:

For the error matrix

$$P_x(k+1) = \left\| \begin{matrix} P_{x11}(k)R^2_{pc}(T) + \sigma^2_{pc}[\ -R^2_{pc}(T)];\ P_{x12}(k)R_{pc}(T)R_{p\pi}(T) + \\ + \sigma_{pc}\sigma_{p\pi}[1 - R_{pc}(T)R_{p\pi}(T)]\,R_{c\pi} \\ P_{x21}(k)\,R_{pc}(T)\,R_{p\pi}(T) + \sigma_{pc}\sigma_{p\pi}[1 - R_{pc}(T)R_{p\pi}(T)]\,R_{c\pi}; \\ P_{x22}(k)\,R^2_{p\pi}(T) + \sigma^2_{p\pi}[1 - R^2_{p\pi}(T)] \end{matrix} \right\|, \tag{1}$$

where $R_{pc}(T) = \exp(-\alpha_{pc}T)$; $R_{r\pi}(T) = \exp(-\alpha_{r\pi}T)$; T is the interval between the measurements; α_{pc} and $\alpha_{r\pi}$ are the parameters of the correlation functions of the actual signal and interference respectively;

For the reciprocal covariation matrix taking into account the apriori reciprocal matrix:

$$P_c(k+1) = \left\| \begin{matrix} \dfrac{[\rho_{22}(T) + \rho_{12}(T) + \sigma_{ш}^2]\,\beta_{11}(k) - [\rho_{11}(T) + \rho_{12}(T)]\,\beta_{12}(k)}{\rho_{11}(T) + 2\rho_{12}(T) + \rho_{22}(T) + \sigma_{ш}^2}; \\ \dfrac{[\rho_{11}(T) + \rho_{12}(T) + \sigma_{ш}^2]\,\beta_{12}(k) - [\rho_{22}(T) + \rho_{11}(T)]\,\beta_{11}(k)}{\rho_{11}(T) + 2\rho_{12}(T) + \rho_{22}(T) + \sigma_{ш}^2}. \\ \dfrac{[\rho_{22}(T) + \rho_{12}(T) + \sigma_{ш}^2]\,\beta_{21}(k) - [\rho_{11}(T) + \rho_{12}(T)]\,\beta_{22}(k)}{\rho_{11}(T) + 2\rho_{12}(T) + \rho_{22}(T) + \sigma_{ш}^2}; \\ \dfrac{[\rho_{11}(T) + \rho_{12}(T) + \sigma_{ш}^2]\,\beta_{22}(k) - [\rho_{22}(T) + \rho_{11}(T)]\,\beta_{21}(k)}{\rho_{11}(T) + 2\rho_{12}(T) + \rho_{22}(T) + \sigma_{ш}^2} \end{matrix} \right\|, \tag{2}$$

where

$$\beta_{11}(k) = P_{11}(k)R_{pc}(T)\Delta R_c(T) + P_{c11}(k)R_{pc}^2(T) + \sigma_{pc}^2[1 - R_{pc}^2(T)];$$
$$\beta_{12}(k) = P_{12}(k)R_{pc}(T)\Delta R_\pi(T) + P_{c12}(k)R_{pc}(T)R_{p\pi}(T) +$$
$$+ \sigma_c\sigma_\pi R_{c\pi}[1 - R_{pc}(T)R_{p\pi}(T)];$$
$$\beta_{21}(k) = P_{21}(k)R_{p\pi}(T)\Delta R_c(T) + P_{c21}(k)R_{pc}(T)R_{p\pi}(T) +$$
$$+ \sigma_c\sigma_\pi R_{c\pi}[1 - R_{pc}(T)R_{p\pi}(T)];$$
$$\beta_{22}(k) = P_{22}(k)R_{p\pi}(T)\Delta R_\pi(T) + P_{c22}(k)R_{p\pi}^2(T) + \sigma_{p\pi}^2[1 - R_{p\pi}^2(T)];$$
$$\Delta R_c(T) = |R_{pc}(T) - R_c(T)|; \qquad \Delta R_\pi(T) = |R_{p\pi}(T) - R_\pi(T)|;$$

For the real covariation error matrix:

$$P_{p\,11}(k+1)=\frac{[\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,11}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}-\frac{[\rho_{11}(T)+\rho_{12}(T)][\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2][P_{p\,21}(k+1/k)+P_{p\,12}(k+1/k)]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{[\rho_{11}(T)+\rho_{12}(T)]^2[P_{p\,22}(k+1/k)+\sigma_{\text{ш}}^2]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}, \tag{3}$$

$$P_{p\,12}(k+1)=\frac{[\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2][\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,12}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}-\frac{[\rho_{11}(T)+\rho_{12}(T)][\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,22}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{-[\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2][\rho_{22}(T)+\rho_{12}(T)]P_{p\,11}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{[\rho_{11}(T)+\rho_{12}(T)][\rho_{22}(T)+\rho_{12}(T)][P_{p\,21}(k+1/k)+\sigma_{\text{ш}}^2]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}, \tag{4}$$

$$P_{p\,21}(k+1)=\frac{[\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2][\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,21}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}-\frac{[\rho_{22}(T)+\rho_{12}(T)][\rho_{22}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,11}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{-[\rho_{11}(T)+\rho_{12}(T)][\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,22}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{[\rho_{22}(T)+\rho_{12}(T)][\rho_{11}(T)+\rho_{12}(T)][P_{p\,12}(k+1/k)+\sigma_{\text{ш}}^2]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2} \tag{5}$$

$$P_{p\,22}(k+1)=\frac{[\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2]P_{p\,22}(k+1/k)}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}-\frac{[\rho_{22}(T)+\rho_{12}(T)][\rho_{11}(T)+\rho_{12}(T)+\sigma_{\text{ш}}^2][P_{p\,12}(k+1/k)+P_{p\,21}(k+1/k)]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}+\frac{[\rho_{22}(T)+\rho_{12}(T)]^2[P_{p\,11}(k+1/k)+\sigma_{\text{ш}}^2]}{\rho_{11}(T)+2\rho_{12}(T)+\rho_{22}(T)+\sigma_{\text{ш}}^2}, \tag{6}$$

where где

$$P_{p11}(k+1/k)=P_{\text{э}11}(k)\Delta R_{\text{с}}^2(T)+2P_{\text{с}11}(k)R_{\text{с}}(T)\Delta R_{\text{с}}(T)+P_{p11}(k)R_{\text{с}}^2(T)+\sigma_{p\text{с}}^2[1-R_{p\text{с}}^2(T)]; \tag{7}$$

$$P_{p12}(k+1/k)=P_{\text{э}12}(k)\Delta R_{\text{с}}(T)\Delta R_{\text{п}}(T)+P_{\text{с}12}(k)R_{\text{п}}(T)\Delta R_{\text{с}}(T)+P_{\text{с}21}(k)R_{\text{с}}(T)\Delta R_{\text{п}}(T)+P_{p12}(k)R_{\text{с}}(T)R_{\text{п}}(T)+\sigma_{p\text{с}}\sigma_{p\text{п}}R_{\text{сп}}[1-R_{p\text{с}}(T)R_{p\text{п}}(T)]; \tag{8}$$

$$P_{p21}(k+1/k)=P_{\text{э}21}(k)\Delta R_{\text{с}}(T)\Delta R_{\text{п}}(T)+P_{\text{с}21}(k)R_{\text{с}}(T)\Delta R_{\text{п}}(T)+P_{\text{с}12}(k)R_{\text{п}}(T)\Delta R_{\text{с}}(T)+P_{p21}(k)R_{\text{с}}(T)R_{\text{п}}(T)+\sigma_{p\text{с}}\sigma_{p\text{п}}R_{\text{сп}}[1-R_{p\text{с}}(T)R_{p\text{п}}(T)]; \tag{9}$$

$$P_{p22}(k+1/k)=P_{\text{э}22}(k)\Delta R_{\text{п}}^2(T)+2P_{\text{с}22}(k)R_{\text{п}}(T)\Delta R_{\text{п}}(T)+P_{p22}(k)R_{\text{п}}^2(T)+\sigma_{p\text{п}}^2[1-R_{p\text{п}}^2(T)]. \tag{10}$$

The results of computer calculations of expressions (3)-(6) for the case where the correlation functions of the model and the actual process of the variation in the level of the interference differ from each other by the correlation interval, i.e., $\alpha_{r\pi} \neq \alpha_{\pi}$, are shown in Figures 1 and 2. In this case, the curves obtained for $\Delta x(k)$ are shown in Figure 1, while those for the absolute value of $\Delta x(k)$ are shown in Figure 2.

It was assumed in the calculations that:

$$\sigma_c^2/\sigma_п^2=10;\quad \xi_{11}(k+1)=P_{p11}(k+1)/\sigma_x^2;\quad \xi_{22}(k+1)=P_{p22}(k+1)/\sigma_x^2;\quad \sigma_x^2=\sigma_c^2+\sigma_п^2+2\sigma_c\sigma_п R_{cп}.$$

As can be seen from the graph shown in Figure 1, when the model of the synthesis of the analyzer-extrapolator for channel quality corresponds to the slow changes in the interference levels ($\alpha_{\pi} = 1/\tau_{kop} = 0.1 = [1/\tau_{corr}]$), the measurement errors increase with a reduction in the correlation interval of the actual process, and when the ratio is $\alpha_{r\pi}/\alpha_{\pi} = 7$, they differ from the presupposed error of the estimate of both the signal and the interference by approximately a factor of five (in this case, it was assumed that $\alpha_{pc} = \alpha_c = 1/\tau_{corr} = 0.1$). When the synthesis model corresponds to the fast changes in the interference level ($\alpha_{r\pi} < \alpha_{\pi}$), the estimate errors increase more slowly.

It can be seen from the functions shown in Figure 2 that the absolute value of the estimate error has a minimum close to the actual value of the interference correlation interval.

An analysis of the results obtained here permit the overall conclusion that to increase the precision of the estimate and the prediction of signal and interference levels in a working channel, in the absence of apriori information on the correlation properties of the interference, it is necessary to adjuct the parameters of the analyzer-extrapolator to match the actual dynamics of the process of the change in the interference level.

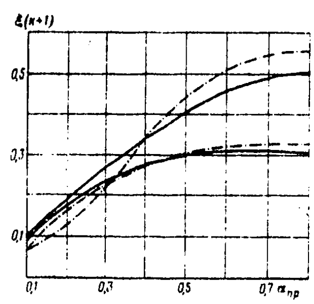

Figure 1. The solid curves are for $\xi_{11}(k+1)$; the dashed and dotted curves are for $\xi_{22}(k+1)$.

$\alpha_c = 0.1$, $R_{c\pi} = 0.3$ and $\sigma_c^2/\sigma_{\pi}^2 = 10$.

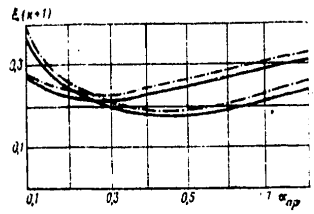

Figure 2. The solid curves are for $\xi_{11}(k+1)$; the dashed and dotted curves are for $\xi_{22}(k+1)$.
$\alpha_c = 0.1$; $R_{c\pi} = 0.3$ and $\sigma_c^2/\sigma_\pi^2 = 10$.

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8225
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ELECTROMAGNETIC WAVE PROPAGATION, ELECTRODYNAMICS

UDC 537.8

ON THE RELATIONSHIP BETWEEN THE ENERGY COEFFICIENTS OF REFLECTION AND THE TRANSMITTANCE OF AN ELECTROMAGNETIC WAVE AT THE BOUNDARY OF TWO ABSORBING MEDIA

Moscow RADIOTEKHNIKA I ELEKTRONIKA in Russian Vol 25, No 7, Jul 80 pp 1524-1526 manuscript received 2 Feb 79

[Paper by M.I. Finkel'shteyn and V.L. Mendel'son]

[Text] The reflection and refraction of a plane electromagnetic wave at the plane boundary of two media, including at the boundary of an absorbing medium has been treated, for example, in [1]. The fundamental theoretical principles for electromagnetic wave propagation in layered media are given in [2]. Interest has recently increased in the relevant applied questions [3]. In a number of cases, it is desirable to use a relationship between the energy coefficents of reflection and transmission to calculate the wave intensities in layered media; this relationship is lacking for the general case of two absorbing media in the known literature.

We shall recall the basic relationships. The average value of the absolute value of the complex Poynting vector for a nonmagnetic medium is written in the form:

$$\Pi = \frac{1}{2}\operatorname{Re}\{\dot{E}\dot{H}^*\} = \frac{1}{2}E^2\operatorname{Re}(1/\dot{W}^*) = \frac{1}{240\pi}E^2\operatorname{Re}\sqrt{\dot{\varepsilon}}, \quad (1)$$

where $\dot{E}$ is the complex amplitude of the electrical field intensity; E is the absolute value; $\dot{H}^*$ and $\dot{W}^*$ are the complex conjugate values of the magnetic field conjugacy [sic] and the characteristic impedance of the medium; $\dot{\varepsilon}$ is the complex relative dielectric permittivity of the medium.

As is well known, the Fresnel coefficients at the boundary of media 1 and 2 are related by the expression:

(2) $$\dot{T}_{1-2} = 1 + \dot{R}_{1-2},$$

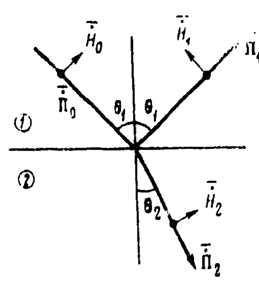
where for the case of vertical incidence at the separation boundary:

(3) $$\dot{R}_{1-2} = \frac{\sqrt{\dot{\varepsilon}_1} - \sqrt{\dot{\varepsilon}_2}}{\sqrt{\dot{\varepsilon}_1} + \sqrt{\dot{\varepsilon}_2}}; \quad \dot{T}_{1-2} = \frac{2\sqrt{\dot{\varepsilon}_1}}{\sqrt{\dot{\varepsilon}_1} + \sqrt{\dot{\varepsilon}_2}}.$$

The energy coefficients of reflection and transmittance for normal incidence at the separation boundary of media 1 and 2 are equal to the following in accordance with (1)

(4) $$R_{p1-2} = \frac{\Pi_1}{\Pi_0} = R^2_{1-2}; \; T_{p1-2} = \frac{\Pi_2}{\Pi_0} = \frac{\mathrm{Re}\sqrt{\dot{\varepsilon}_2}}{\mathrm{Re}\sqrt{\dot{\varepsilon}_1}} T^2_{1-2}.$$

where Π_0. Π_1 and Π_2 are the power flux densities for the incident, reflected and refracted waves respectively.



To determine the relationship between the coefficients $R_{p\ 1-2}$ and $T_{p\ 1-2}$ for non-absorbing media, we start on the basis of the law governing the conservation of energy from the noncontinuity of the normal component of the overall power flux density when passing through the boundary [1], i.e.:

(5) $$\bar{n}(\bar{\Pi}_0 + \bar{\Pi}_1) = \bar{n}\bar{\Pi}_2$$

($\bar{n}$ is a unit vector normal to the separation boundary), from which:

(6) $$R_{p\,1-2} + T_{p\,1-2} = 1.$$

Along with this, for absorbing media where two opposing waves are present, the total energy flux in the general case is not equal to the sum of the energy fluxes of the partial waves. What has been said can be seen from expression (6), if it is assumed that the dielectric permittivities of media 1 and 2 are equal to $\dot{\varepsilon}_{1,2} = \varepsilon_{1,2}e^{-j\psi_{1,2}}$, i.e.:

$$\operatorname{Re}\sqrt{\dot\varepsilon_{1,2}}=\sqrt{\varepsilon_{1,2}}\cos(\psi_{1,2}/2);\quad \operatorname{Im}\sqrt{\dot\varepsilon_{1,2}}=\sqrt{\varepsilon_{1,2}}\sin(\psi_{1,2}/2).$$

By substituting in (3), we obtain the following after simple transformations taking (4) into account:

$$R_{p1-2}+T_{p1-2}=\frac{\left(\varepsilon_1+\varepsilon_2-2\sqrt{\varepsilon_1\varepsilon_2}\cos\frac{\psi_1-\psi_2}{2}\right)\cos(\psi_1/2)+4\sqrt{\varepsilon_1\varepsilon_2}\cos(\psi_2/2)}{\left(\varepsilon_1+\varepsilon_2+2\sqrt{\varepsilon_1\varepsilon_2}\cos\frac{\psi_1-\psi_2}{2}\right)\cos(\psi_1/2)},\tag{7}$$

i.e., the equality (6) is justified when $\psi_1 = 0$ and for an arbitrary ψ_2, but is not observed when $\psi_1 \neq 0$ and $\psi_2 = 0$ (when a wave makes a transition from a lossy medium to a lossless medium), as well as in the general case of two lossy media ($\psi_1 \neq 0$, $\psi_2 \neq 0$).

To obtain a universal relationship between $R_{p\,1-2}$ and $T_{p\,1-2}$ within the framework of the energy relationships, it is necessary to proceed from the fact that when transiting through a media boundary, the continuity of the normal component is observed not for the overall energy flux [5], but for the energy flux of the total field [3]. Thus, for the case of oblique incidence, for example, in the case of horizontal polarization (see the Figure), we obtain in accordance with (1):

$$\operatorname{Re}\{(\dot E_0+\dot E_1)[(\dot H_0\cos\Theta_1)^*+(\dot H_1\cos\Theta_1)^*)\}=\operatorname{Re}\{\dot E_2(\dot H_2\cos\Theta_2)^*\},\tag{8}$$

where $\dot E_0$ and $\dot H_0$ are the complex amplitudes of the intensities of the incident wave.

Since $\dot R_{1-2} = \dot E_1/\dot E_0$ then:

$$\dot H_1^*=-\frac{1}{120\pi}(\sqrt{\dot\varepsilon_1})^*\dot E_0^*\dot R_{1-2}^*;\quad \dot H_0^*=\frac{1}{120\pi}(\sqrt{\dot\varepsilon_1})^*\dot E_0^*.\tag{9}$$

Additionally, the power transmittance coefficient for the case of oblique incidence is:

$$T_{p1-2}=\frac{\bar n\bar\Pi_2}{\bar n\bar\Pi_0}=\frac{\operatorname{Re}[\dot E_2(\dot H_2\cos\Theta_2)^*]}{\operatorname{Re}[\dot E_0(\dot H_0\cos\Theta_1)^*]}=\frac{\operatorname{Re}(\sqrt{\dot\varepsilon_2}\cos\Theta_2)^*}{\operatorname{Re}(\sqrt{\dot\varepsilon_1}\cos\Theta_1)^*}T_{1-2}^2.\tag{10}$$

By using the following representations:

$$\dot R_{1-2}=R_{1-2}e^{-j\varphi_{1-2}};\quad (\sqrt{\dot\varepsilon_1})^*=\operatorname{Re}\sqrt{\dot\varepsilon_1}+j\operatorname{Im}\sqrt{\dot\varepsilon_1},$$

following the substitution of (9) in (8) and taking expressions (10) into account, as well as $R_{p\,1-2} = R_{1-2}^2$ (see (4)), we obtain:

(11)
$$R_{p1-2} + T_{p1-2} = 1 + 2R_{1-2}\frac{\operatorname{Im}\sqrt{\dot{\varepsilon}_1}}{\operatorname{Re}\sqrt{\dot{\varepsilon}_1}}\sin\varphi_{1-2} =$$

$$= 1 + 4\operatorname{Im}\sqrt{\dot{\varepsilon}_1}\,\frac{\operatorname{Im}\sqrt{\dot{\varepsilon}_2} - \dfrac{\operatorname{Re}\sqrt{\dot{\varepsilon}_2}}{\operatorname{Re}\sqrt{\dot{\varepsilon}_1}}\operatorname{Im}\sqrt{\dot{\varepsilon}_1}}{|\sqrt{\dot{\varepsilon}_1}|^2 + |\sqrt{\dot{\varepsilon}_2}|^2 + 2\operatorname{Re}\sqrt{\dot{\varepsilon}_1}\operatorname{Re}\sqrt{\dot{\varepsilon}_2} + 2\operatorname{Im}\sqrt{\dot{\varepsilon}_1}\operatorname{Im}\sqrt{\dot{\varepsilon}_2}}.$$

The same relationship obtains for vertical polarization. It changes to (6) in the case where medium 1 is nonabsorbing, i.e., its dielectric permittivity is a real quantity.

In conclusion, for a comparison with (6), we shall compute (11) when a radio wave transits from sea ice into the air. For a frequency range of 10 to 100 MHz, in accordance with [4], $\dot{\varepsilon}$ changes from 12.3 - j17.2 to 8.1 - j7.7, from which $R_{p\,1-2} + T_{p\,1-2} \simeq 1.1$.

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8225
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INSTRUMENTS, MEASURING DEVICES AND TESTERS, METHODS OF MEASURING, GENERAL EXPERIMENTAL TECHNIQUES

UDC 535.854

PROCESSING OF SIGNALS OF A HETERODYNE INTERFEROMETER BY MEANS OF AN AUTOMATIC PHASE CONTROL SYSTEM

Leningrad IVUZ, PRIBOROSTROYENIYE in Russian Vol 23, No 7, 1980 pp 72-75
manuscript received 18 Apr 79

[Article by A.V. Koryabin and V.I. Shmal'gauzen, Moscow State University imeni M.V. Lomonosov]

[Text] The influence of external effects on the signal spectrum of a heterodyne interferometer is discussed. A demonstration is given of the effectiveness of employing automatic phase control for processing the signal.

Heterodyne (two-frequency) interferometers are used at the present time for acoustical measurements [1, 2]. In [1] an expression was arrived at for the sensitivity of an instrument of this sort when taking into account the noise of the electronic equipment. It is known that forms of external instability (thermal, vibrational, etc.) represent one of the main sources of errors in single-frequency interferometers [3]. Therefore, it is interesting to discuss the influence of external effects on the performance of heterodyne systems.

Let us write the alternating-current component of the voltage in the heterodyne interferometer's photodetector (fig 1) in the form [2]:

$$u(t)=U_0\cos\left\{\omega_0 t+\frac{4\pi}{\lambda}[L(t)+x(t)]\right\}+n(t),\qquad U_0=\text{const}, \tag{1}$$

where ω_0 is the heterodyning frequency; λ is the wavelength of the light employed; $n(t)$ is the fluctuation component caused by the photodetector's shot noise; $L(t)$ describes the random motion of the system caused by external effects; and $x(t)$ is determined by the studied motion of the system and for slight harmonic vibrations with an amplitude of a and a cycle frequency of Ω equals

$$x(t)=a\cos\Omega t,\ \ a\ll\lambda. \tag{2}$$

In this case, utilizing the properties of Bessel functions, for $u(t)$ we arrive at the expression

$$u(t)\simeq A_0\cos[\omega_0 t+\Phi]-A_1\cos[\omega_1 t\pm\Phi]-A_{-1}\cos[\omega_{-1}t+\Phi]+n(t), \tag{3}$$

where

$$A_0=U_0J_0(b)\simeq U_0;\ A_1=A_{-1}=U_0J_1(b)\simeq U_0\cdot\frac{b}{2};\ \Phi(t)=\frac{4\pi}{\lambda}\,L(t);\ \omega_{\pm 1}= =\omega_0\pm\Omega;$$

and $J_k(b)$ is a Bessel function of the k-th order of argument $b = 4\pi a/\lambda$.

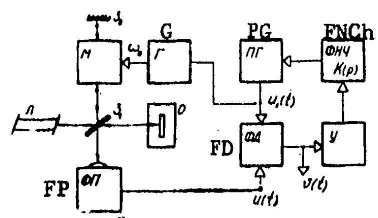

Figure 1. Structural Diagram of Heterodyne Interferometer: L--light source; Z_1--light divider; Z_2--reference arm reflector; O--system studied; M--electrooptical light modulator; G--modulator's heterodyne; FP--photodetector; FD--FAPCh [automatic phase control (APC)] phase detector; U--error signal amplifier; FNCh--low-frequency filter with transmission coefficient of K(p) ; PG--tunable (controlled) FAPCh oscillator

The spectrum of external effects usually has a maximum at low frequencies [3]; therefore, it is possible to use as a mathematical model, $\Phi(t)$, the normal stationary process whose correlation function and spectral density are determined by the variance, σ_0 , and width of the spectrum, α :

$$k(\tau)=\sigma_0^2e^{-\alpha|\tau|},\ \ S(\omega)=2\sigma_0^2\,\frac{\alpha}{\alpha^2+\omega^2}. \tag{4}$$

Assuming n(t) to be normal white noise with an intensity of N_0, we write the spectral density, S_u, of signal (1) in the following manner:

$$S_u(\omega) = \sum_{l=-1}^{1} [S_l(\omega + \omega_l) + S_l(\omega - \omega_l)] + \frac{N_0}{2}, \quad \omega_l = \omega_0 + l\Omega, \tag{5}$$

where $S_l(\omega \pm \omega_l)$ for the model selected for process $\Phi(t)$ equals [4]

$$S_l(\omega \pm \omega_l) = \frac{A_l^2}{2\alpha} e^{-\sigma_0^2} \left[\pi\delta\left(\frac{\omega \pm \omega_l}{\alpha}\right) + F(\omega \pm \omega_l)\right],$$
$$F(\omega \pm \omega_l) = \sum_{n=1}^{\infty} \frac{\sigma_0^{2n}}{n!\,n} \left[1 + \left(\frac{\omega \pm \omega_l}{\alpha n}\right)^2\right]^{-1}. \tag{6}$$

From equation (6) it follows that the influence of instabilities, $\Phi(t)$, on the spectrum of signal (1) consists in a reduction of its strength, determined by factor $\exp(-\sigma_0^2)$ and in the appearance in the region of frequency ω_1 of a "pedestal" described by function $F(\omega \pm \omega_1)$, whose width increases with an increase in σ_0^2 and α (fig 2).

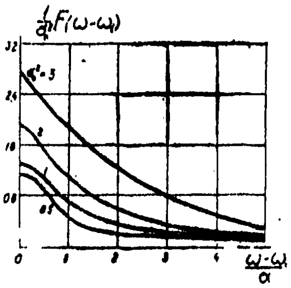

Figure 2. Dependence of "Pedestal" on Variance, σ_0^2, and Width of Spectrum, α

It is possible to reduce the influence of external effects by using an automatic phase control (FAPCh) system, which would result in slow fluctuation, $\phi(t)$, of the phase of signal (1). In the synchronous mode the FAPCh controlled oscillator produces the signal

$$u_1(t) = U_1 \sin[\omega_0 t + \Phi_1(t)];$$
$$\Phi_1(t) \simeq \Phi(t), \tag{7}$$

and the voltage at the output of the FAPCh phase detector equals [5]

$$v(t) = \frac{1}{2}\mu U_0 U_1 \sin\left[\varepsilon(t) - \frac{4\pi}{\lambda}x(t)\right], \quad \varepsilon = \Phi_1 - \Phi; \tag{8}$$

where μ is the phase detector's transmission coefficient. Assuming the transmission coefficient, K(p) , of the low-frequency filter (cf. fig 1) in the noise frequency band to be equal to one (first-order FAPCh system) and assuming the variance, σ_ε^2 , of the tracking error, ε , to be sufficiently slight [5], for (8) we get

$$v(t) \simeq V_0\varepsilon(t) - y(t), \quad V_0 = \frac{1}{2}\mu U_0 U_1 J_0(b), \tag{9}$$

where

$$y(t) = V_1 \cos[\Omega t + \varepsilon(t)] + V_1 \cos[\Omega t - \varepsilon(t)],$$
$$V_1 \simeq V_0 \frac{b}{2}. \tag{10}$$

The spectrum of the signal for error ε , taking into account (4), can be written in the form [5]:

$$S_\varepsilon(\omega) = \frac{2\sigma_0^2 \cdot \alpha \cdot \omega^2}{(\alpha^2 + \omega^2)(\beta^2 + \omega^2)} + \frac{\frac{N_2}{2} + \frac{1}{2q}}{\beta^2 + \omega^2}, \tag{11}$$

where β is the width of the FAPCh holding band, N_2 is the intensity of fluctuations of the phase of the tuned oscillator, and q is the ratio of the power of carrier (1) to the intensity of shot noise in the frequency band, β ,

$$q = \frac{U_0^2}{N_0 2\beta}. \tag{12}$$

The variance of process $\varepsilon(t)$ is determined by the equation

$$\sigma_\varepsilon^2 = \sigma_0^2 \frac{\alpha}{\alpha+\beta} + \frac{1}{4}\left(\frac{1}{q} + \frac{N_2}{\beta}\right). \tag{13}$$

If the following conditions are fulfilled:

$$\beta \gg \alpha,\ \beta \gg N_2,\ q \gg 1, \tag{14}$$

then $\sigma_\varepsilon^2 \leq 1$ and the correlation function of process $y(t)$ can be written in the form [4]:

$$k_y(\tau) \simeq V_1^2 e^{-\sigma_\varepsilon^2} [1 + k_\varepsilon(\tau)] \cos \Omega\tau, \tag{15}$$

where $k_\varepsilon(\tau) = C_1 e^{-\beta|\tau|} - C_2 e^{-\alpha|\tau|}$;

$$C_1 = \sigma_0^2 \frac{\alpha\beta}{\beta^2 - \alpha^2} + \frac{1}{4}\left(\frac{1}{q} + \frac{N_2}{\beta}\right); \quad C_2 = \sigma_0^2 \frac{\alpha^2}{\beta^2 - \alpha^2}. \tag{16}$$

From (15) and (16) by means of a Fourier transform we get an expression for the spectral density of signal $v(t)$:

$$S_v(\omega) = V_0^2 S_\varepsilon(\omega) + S_y(\omega+\Omega) + S_y(\omega-\Omega), \tag{17}$$

$$\begin{cases} S_y(\omega \pm \Omega) = \dfrac{V_1^2}{\alpha} e^{-\sigma_\varepsilon^2}\left[\pi\delta\left(\dfrac{\omega \pm \Omega}{\alpha}\right) + F_1(\omega \pm \Omega)\right], \\ F_1(\omega \pm \Omega) = \dfrac{\alpha}{\beta} C_1\left[1 + \left(\dfrac{\omega \pm \Omega}{\beta}\right)^2\right]^{-1} - C_2\left[1 + \left(\dfrac{\omega \pm \Omega}{\alpha}\right)^2\right]^{-1}. \end{cases} \tag{18}$$

A comparison of (6) and (18) demonstrates that the influence of FAPCh on the spectrum of the signal processed consists in lowering of the "pedestal" near frequency Ω and in a change in its shape (fig 3); the use of FAPCh also considerably reduces the dependence of the spectrum on the statistical characteristics, (4), of the external effect, since upon fulfillment of the conditions in (14) the value of $\exp(-\sigma_\varepsilon^2) \simeq 1 - \sigma_\varepsilon^2$ depends but slightly on the variance and width of the noise spectrum.

The stabilizing effect of FAPCh was verified in a heterodyne interferometer with an electrooptical light modulator similar to that described in [6] in measuring the amplitude of the ultrasonic vibrations of a piezoceramic plate. The apparatus had the following parameters: $\omega_0/2\pi = 1$ MHz , $\beta/2\pi = 20$ kHz and $N_2/2\pi = 0.5$ kHz . When the FAPCh was shut off the second input of the phase detector was connected to the heterodyne of the electrooptical unit (cf. fig 1), which had a stability of $\sim 10^{-6}$. The spectra of the output voltage of the phase detector (fig 4) obtained by means of an S4-12 spectrum analyzer confirm the effectiveness of using an FAPCh system for processing the signals of a heterodyne interferometer.

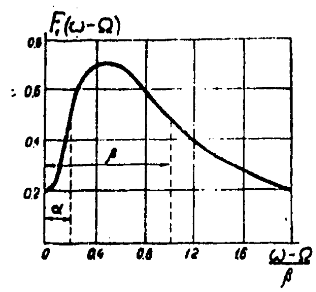

Figure 3. Spectrum of Signal at Output of FAPCh System Near Frequency Ω ; $C_2 = 0.8 \cdot C_1(\alpha/\beta)$

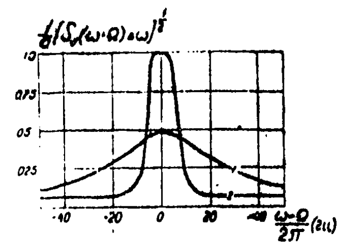

Figure 4. Spectrum of Vibration $a \cos \Omega t$ with the FAPCh System Turned Off, 1, and On, 2: $a = 1.2$ Å , $\Omega/2\pi = 70$ kHz , $\alpha/2\pi = 30$ Hz , $\sigma_0 \simeq \pi$, and $\Delta\omega/2\pi = 7$ Hz is the transmission band of the spectrum analyzer

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PUBLICATIONS, INCLUDING COLLECTIONS OF ABSTRACTS

UDC 621.38:537.312.5

BOOK CONCERNED WITH EFFECT OF PENETRATING RADIATION ON ELECTRONIC EQUIPMENT

Moscow DEYSTVIYE PRONIKAYUSHCHEY RADIATSII NA IZDELIYA ELEKTRONNOY TEKHNIKI [Effect of Penetrating Radiation on Electronic Equipment] in Russian 1980 signed to press 17 Mar 80 pp 2, 224

[Annotation and table of contents from book by Valeriy Mikhaylovich Kulakov, Yevgeniy Aleksandrovich Ladygin, Valeriy Ivanovich Shakhovtsov, Erikh Nikolayevich Vologdin and Yuriy Nikolayevich Andreyev, Izdatel'stvo "Sovetskoye radio," 6500 copies, 224 pages]

[Text] The results of domestic and foreign study of the effect of radiation on variation pieces of electronic equipment are generalized and set forth in the book. The physical mechanisms of radiation damage and the nature and properties of radiation flaws in materials and pieces of electronic equipment are examined. Basic mathematical relationships and a large number of experimentally derived dependencies for the change in the parameters of the primary classes of electronic equipment and materials when exposed to neutrons, protons, electrons and Gamma-quanta are given.

The book is intended for specialists engaged in the design of electronic equipment and research into their radiation stability, and it may be useful for instructors and students in higher educational institutions.

Contents

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BOOK CONCERNED WITH MICROCIRCUIT DESIGN AND TECHNOLOGY

Moscow KONSTRUKTSIYA I TEKHNOLOGIYA MIKROSKHEM (GIS I BGIS) [Microcircuit Design and Technology (Hybrid IC and Large Hybrid IC)] in Russian 1980 signed to press 9 May 1979 pp 2, 253-254

[Annotation and table of contents from book by Yuriy Petrovich Yermolayev, Mikhail Fedorovich Ponomarev and Yuriy Grigor'yevich Kryukov, Izdatel'stvo "Sovetskoye radio," 25,000 copies, 256 pages]

[Text] Methods for the design of passive film elements and methods for their manufacture are presented. Means for optimization of the topology of sets of film components are examined from the positions of stability and accuracy of the microcircuits' output parameters. General questions about the designing of hybrid integrated microcircuits (thermal conditions, spurious couplings, reliability) are set out. Assembly and mounting procedures, as well as organization of the production of hybrid integrated microcircuits (HIC) and large hybrid integrated microcircuits (LHIC) are described. Attention is devoted to microcircuits for the microwave band and to functional microelectronic components and apparatus.

(The text) is intended for students studying in the areas of specialization: "Design and Production of Radio Equipment" and "Design and Production of Electronic Computer Devices."

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[31-9194]

9194
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BOOK CONCERNED WITH PANORAMIC RECEIVERS AND SPECTRUM ANALYZERS

Moscow PANORAMNYYE PRIYEMNIKI I ANALIZATORY SPEKTRA [Panoramic Receivers and Spectrum Analyzers] in Russian 1980 signed to press 25 Dec 79 pp 2, 348-350

[Annotation and table of contents of book by Valentin Alekseyevich Martynov and Yuriy Ivanovich Selikhov, Izdatel'stvo "Sovetskoye radio," 10,000 copies, 350 pages]

[Text] The basic methods of spectral analysis and the principles for construction of panoramic receivers and spectrum analyzers are examined. The designation and areas of application for panoramic devices are determined, and requirements for their basic characteristics are formalized. Physical processes in the primary channels of these devices are analyzed and recommendations for their optimum construction under real operating conditions are made. Structural circuits of panoramic devices and variants in their construction are examined, as are questions concerning the display, registration and processing of the results. Descriptions of certain panoramic receivers and spectrum analyzers presently in use are given (the first edition came out in 1964).

(The book) is intended for specialists engaged in the development of pertinent equipment or who use it for analysis of radio radiation. It may be of use to instructors and students of higher educational institutions for the corresponding specialties.

Contents

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BOOK CONCERNED WITH PROBLEMS OF THE QUALITY OF RADIO COMPONENTS

Moscow VOPROSY KACHESTVA RADIOETALEY [Problems of the Quality of Radio Components] in Russian 1980 signed to press 25 Feb 80 pp 2, 327-328

[Annotation and table of contents from book by Boris Yuzikovich Gelikman, Galina Anatol'yevna Goryacheva, Lev Leonidovich Kristalinskiy and Vladimir Vladimirovich Stal'bovskiy, Izdatel'stvo "Sovetskoye Radio," 9000 copies, 328 pages]

[Text] Problems are covered concerned with providing for and controlling the quality of capacitors and resistors under conditions of mass production and use. Particular attention is paid to an analysis of the means of insuring high quality and reliability of radio components, to the physics of capacitor and resistor failures and to examining existing methods of monitoring the quality of these items.

(The book) is intended for specialists working at the design and production of radio components and electronic equipment.

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RADARS, RADIONAVIGATION AIDS, DIRECTION FINDING, GYROS

UDC 621.396.96

SUBSURFACE RADAR SOUNDING OF LAYERED MEDIA. THE EQUATION OF THE SOUNDING BEAM

Moscow RADIOTEKHNIKA I ELEKTRONIKA in Russian Vol 25, No 7, Jul 80 pp 1434-1437 manuscript received 31 Jan 79

[Paper by V.S. Luchininov]

[Text] The sounding ray path in a layered medium is defined as the shortest range path via which the sounding signal propagates from the transmitter to the target and back. An equation is derived for the sounding path in an observed medium and the possibilities of refining the procedure for subsurface sounding and vertical radar profiling through introducing a doppler frequency shift meter into the circuitry of the radar system are analyzed.

Let a two position radar system (RLS) probing with nondirectional antennas travel above an inhomogeneous half-space of a layered structure.

The relationship between the spatial position of the incident and reflected beams as well as the signal parameters and the motion of the radar system and the target are of interest for the practice of subsurface sounding of layered media for the purpose of reconstructing the relief of the structural boundaries or determining the coordinates of concentrated targets.

We shall find this relationship given the following limitations:
1) The ranges from the radar transmitter and receiver to the boundaries of the nearest layer, as well as the characteristic dimensions of the layers are much greater than the wavelength, corresponding to the lowest frequency in the spectrum of the sounding signal in each layer.

2) The boundaries of the layers are stationary, isotropic, while their deviation from the corresponding mean values and their slopes are small; 3) The components of the structure are weakly absorbing media; 4) The motion of the radar transmitter and receiver as well as the target are mutually independent, while their displacement over the signal propagation time from the transmitter to the receiver are neglectibly small.

The system of limitations enumerated above produces a satisfactory approximation on one hand for the conditions for radar photography of glaciological objects and geological structures from space vehicles and aircraft (KLA and ALA respectively) [1, 2, 3], and on the other hand, substantiate the applicability of the geometric optics approximation.

We shall number the sequential layers in which the sounding signal propagates, beginning with the layer in which the radar is located, with the numbers 1, 2, 3, ..., m, and let m be the number of the reflecting surface or medium in which a concentrated target is located. Further, we shall designate the boundary surfaces of the media as S_j, where $j = 1, 2, 3, \ldots, m$, while the refraction or reflection points of the sounding beam (ZL) of the surface S_m at these boundaries, respectively as Q_j^t, Q_m, Q_j^r, where the superscripts designate the segments of the sounding beam from the transmitter to the reflector (t) and from the reflector to the receiver (r). We shall designate the radar source and receiver as Q_0^t and Q_0^r respectively.

With the symbols introduced here, the optical length of the sounding beam segment in the layer (j) can be represented in the form:

$$\mathcal{L}_j^t = n_j Q_{j-1}^t Q_j^t, \tag{1}$$

where n_j is the absolute index of refraction of the material of the layer (j).

We rewrite (1) in the form:

$$\mathcal{L}_j^t = n_j \sqrt{(X_{j-1}-X_j)^2+(Y_{j-1}-Y_j)^2+(Z_{j-1}-Z_j)^2}, \tag{2}$$

where X_{j-1}^t, Y_{j-1}^t, Z_{j-1}^t and X_j^t, Y_j^t and Z_j^t are the coordinates of the points Q_{j-1}^t and Q_j^t respectively, and we differentiate the expression obtained with respect to time.

Taking into account the fact that $\dot{n}_j = 0$, we obtain:

(3) $$\dot{\mathscr{L}}_j^t = \vec{S}_j^t\{\vec{V}(Q_j^t) - \vec{V}(Q_{j-1}^t)\},$$

where $\vec{S}_j^t$ is the beam vector in the layer (j); $\vec{V}(Q_j^t)$ is the velocity of the point Q_j^t.

Let the radar target be the structural surface S_m.

By summing expressions of the form of (3) over all segments of the sounding beam from the source to the reflector and from the reflector to the receiver, we obtain:

(4)
$$\begin{aligned}\sum_{j=1}^{m} (\dot{\mathscr{L}}_j^t + \dot{\mathscr{L}}_j^r) = {} & [\vec{S}_1^r \vec{V}(Q_0^r) - \vec{S}_1^t \vec{V}(Q_0^t)] + \\ & + \sum_{j=1}^{m-1} \vec{V}_j(Q_j^t)(\vec{S}_j^t - \vec{S}_{j+1}^t) + [\vec{S}_m^t \vec{V}(Q_m^t) - \vec{S}_m^r \vec{V}(Q_m^r)] + \\ & + \sum_{j=m-1}^{1} \vec{V}(Q_j^r)(\vec{S}_{j+1}^r - \vec{S}_j^r).\end{aligned}$$

We shall show that on the right side of the expression written here, all of the terms with the exception of the first are equal to zero.

We shall consider an expression of the form:

$$\vec{V}(Q_j^t)(\vec{S}_j^t - \vec{S}_{j+1}^t),$$

which takes the form of the scalar product of the velocity vector of the refraction point Q_j^t at the separation surface S_j, and the vector equal to the difference in the beam vectors in the contiguous layers.

The refraction point Q_j^t is invariably related to the separation surface S_j, and therefore its velocity vector is located in a plane tangent to the separation surface at the same point.

The vector $(\vec{S}_j^t - \vec{S}_{j+1}^t)$, by virtue of the law of refraction, agrees with the vector of the normal to the surface at the refraction point within the precision of the scalar factor [4].

Consequently, the vectors $\vec{V}(Q_j^t)$ and $(\vec{S}_j^t - \vec{S}_{j+1}^t)$ are orthogonal, and their scalar product is equal to zero.

Thus, any term of each of the sums incorporated in the right side of expression (4) is equal to zero.

As far as the following term is concerned:

$$[\vec{S}_m{}^t\vec{V}(Q_m{}^t)-\vec{S}_m{}^r\vec{V}(Q_m{}^r)],$$

then the points Q_m^t and Q_m^r coincide, it proves to a special case of the expression already considered above and for the same reasons should be set equal to zero.

Finally, we obtain instead of (4):

(5) $$c\dot{\tau}_m=\vec{S}_1{}^r\vec{V}(Q_0{}^r)-\vec{S}_1{}^t\vec{V}(Q_0{}^t),$$

where c is the velocity of light in a vacuum; $\tau_m=(1/c)\sum_{j=1}^{m}(\mathscr{L}_j{}^t+\mathscr{L}_j{}^r)$ is the delay of the reflected signal; $\dot{\tau}_m$ is its derivative with respect to time.

In a similar fashion, an equation can be derived for a concentrated target in the layer (m):

(6) $$c\dot{\tau}_m=\vec{S}_1{}^r\vec{V}(Q_0{}^r)-\vec{S}_1{}^t\vec{V}(Q_0{}^t)+\vec{V}(Q_m)(\vec{S}_m{}^t-\vec{S}_m{}^r),$$

where the quantities τ_m and c have the same meaning as before. The expressions obtained remain applicable when the source and the receiver are located in different layers or in a continuously inhomogeneous medium.

In the case of a single position radar, we have:

a) For a reflecting surface

(7) $$\vec{S}_1{}^t\vec{V}(Q_0{}^t)=-c\dot{\tau}_m/2,$$

(b) For a concentrated target

(8) $$\vec{S}_m{}^t\vec{V}(Q_m)-\vec{S}_1{}^t\vec{V}(Q_0{}^t)=c\dot{\tau}_m/2.$$

If the target is stationary, equation (8) changes to (7).

Let P be a fixed point in space, through which identical radar soundings are made at different points in time and at different velocities.

If the above mentioned velocities form a base in three-dimensional space, the equation for the sounding beam of the surface S_m in the observed medium can be written in the form:

$$\frac{X_1V_1}{\dot{\tau}_{1m}}=\frac{X_2V_2}{\dot{\tau}_{2m}}=\frac{X_3V_3}{\dot{\tau}_{3m}}, \tag{9}$$

where X_i (i = 1, 2, 3) are the projections of the sounding beam onto the corresponding coordinate axes; V_i are the absolute values of the velocities, where the origin of the coordinates cooncides with the point P.

Expression (9) also permits other forms of writing, for example:

$$\frac{X_1V_1}{\Omega_1}=\frac{X_2V_2}{\Omega_2}=\frac{X_3V_3}{\Omega_3}, \tag{10}$$

where Ω_i is the doppler frequency corresponding to the vector $\vec{V}_i$, or

$$\frac{X_1}{\left(\frac{d\tau_{1m}}{dx_1}\right)}=\frac{X_2}{\left(\frac{d\tau_{2m}}{dx_2}\right)}=\frac{X_3}{\left(\frac{d\tau_{3m}}{dx_3}\right)}, \tag{11}$$

where the spatial derivatives of the delay for the case of motion along the axes of some baseline are designated as $(d\tau_{im}/dx_i)$.

As a rule, the observer is located invariably above the surface of the object being sounded, something which makes it possible to employ a plane base and an abreviated form for the writing of the sounding path equation.

Time scans (radargrams), which fix the instantaneous distributions of the delays of the reflected signals relative to the sounding signal for the case of motion along one of the axes of the velocity baseline being imaged [1, 2, 3] are employed in radar systems which are actually used for sounding to reconstruct the relief of the structural boundaries of a medium.

This creates difficulties in the decoding of the radargrams of layered structures with a complex relief of the reflecting boundaries [3, 5].

The analysis made here indicates the possibility of refining the procedure of both subsurface sounding and vertical profiling through the introduction of copple frequency meters into the radar circuitry and the creation of a velocity baseline at each point in the trajectory of a space vehicle or aircraft, with respect to which the direction of the sounding beam can be determined.

In actual fact, by having data on the radar coordinates, the elements of the velocity base, the doppler frequencies of the reflected signals and the propagation velocities of the radio waves in the layers, one can determine the position of the sounding beam in the observed medium by means of equation (10), and then, based on the known delay, determine the coordinates of the points of the reflecting boundaries beneath the radar.

The procedure described here is always feasible for an open reflecting surface, S_1, since in the case of a single position experiment, a single segment of the sounding beam coincides with the normal to the surface at the reflection point.

To obtain the coordinates of a subradar point on a high numbered surface, for example, S_m, it is necessary to supplement the data enumerated above with equations of surfaces having smaller numbers from the first to the (m - 1)-th inclusively.

The coordinates of the refraction points, the sounding beam directions and the partial delays for the case of propagation in media with numbers of from 1 to m are sequentially computed from these data, after which the coordinates of the reflection point Q_m are determined.

The additional data can be obtained from the results of detailed plane surveying.

In the case of a profile survey, the information on the direction of the sounding beam can be used to eliminate signals from lateral directions.

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ON THE EFFECTIVENESS OF INTERPERIOD COMPENSATION FOR PASSIVE RADAR JAMMING

Moscow RADIOTEKHNIKA I ELEKTRONIKA in Russian Vol 25, No 7, Jul 80 pp 1534-1537 manuscript received 23 Jan 79

[Paper by V.A. Kortunov and F.V. Kivva]

[Text] Moving target indication (SDTs) [MTI] filters have found application in the detection and measurement microwave signals against a background of returns from local objects, vegetative cover, an agitated sea surface and weather formations. In one of the most widespread variants - an interperiod compensation (ChPK) [IPC] filter, the frequency characteristic is described by the formula:

$$H_1(\omega) = 2^{2p} \sin^{2p} \frac{\omega}{2F_п}, \tag{1}$$

where $F_П$ is the repetition rate of the sounding signals; p is the subtraction multiplicity factor (p = 1, 2, ...).

The increase in the signal-to-noise ratio is achieved by the predominant suppression of the narrow band interference (compared to the signal), where the bulk of the interference energy is concentrated close to the harmonics of the repetition rate. It has been established [1] that with a gaussian envelope of the power spectrum of the interference, a characteristic of the improvement of the signal to noise ratio is the increasing function of the subtraction multiplicity factor.

Theoretical [3] and experimental [4] studies indicate the possibility of approximating the power spectrum of the interference with an exponential function

(2)
$$G_{N2}(\omega)=\frac{G_{N2}(\omega_1)}{1+\left|\frac{\omega-\omega_1}{\Omega}\right|^{\nu}}$$

where $\Omega = 2\pi F_e$; ν is the exponent which is determined experimentally. In particular, for the case of returns from forests, it varies in a range of -3 to -4.8 with the most probable value being -3.8 [4]. In contrast to a gaussian curve, the approximation of the curve (2) of the measured spectra of returns from the sea in the 1 and 30 GHz bands proves to be true all the way down to levels of -25 to -30 dB and lower for values of the parameter ν which vary in a range of 2.5 to 4.

The purpose of this paper is to calculate the effectiveness of multiple ChPK [interperiod compensation] circuits for the case of the approximation of the interference spectrum with an exponential function. Results calculated using a general procedure for a gaussian approximation of the interference spectrum are also given.

Improving the Signal-to-Noise Ratio By Means of p-Multiple Interperiod Compensation

The selective properties of IPC can be estimated in various ways. However, most frequently used is the calculation of the subnoise visibility coefficient, defined as the ratio of the interference power at the input to the suppression circuit to that useful signal power at the input for which the average power of the useful signal is nine times greater than the interference power at the output of the circuit [2]:

(3)
$$\varepsilon=\frac{N_{\text{вх}}/S_{\text{вх}}}{9N_{\text{вых}}/S_{\text{вых}}}=\frac{K}{9}\frac{N_{\text{вх}}}{N_{\text{вых}}} = \frac{K}{9}\,\frac{N_{in}}{N_{out}}$$

where N and S are the powers of the interference and the signal, the subscripts "in" and "out" apply to the input and output terminals of the circuit; $K = \langle S_{out}\rangle/S_{in}$ is the ratio of the useful signal power averaged over the period of the frequency characteristic of the IPC at the output to the power of the useful signal at the input to the IPC:

(4)
$$K=\frac{1}{2\pi F_{\text{п}}}\int_0^{2\pi F_{\text{п}}} H(\omega)\,d\omega=\int_0^1 \sin^{2p}\left(\pi\frac{f}{F_{\text{п}}}\right) d\left(\frac{f}{F_{\text{п}}}\right);$$

$H(\omega) = H_1(\omega)/H_{max}(\omega)$ is the normalized frequency characteristic of the p-multiple IPC.

Shown in the Table are the values of the coefficients K for IPC's with multiplicity factors of from 1 to 5, computed from formula (4).

TABLE

p	1	2	3	4	5
K	0,5	0,375	0,314	0,27	0,24

It can be seen from the Table that with an increase in the value of p, the average signal power at the output of the IPC circuit decreases. For this reason, the question of the expediency of increasing the multiplicity factor can be resolved only by taking into account the reduction in the interference power at the output.

By normalizing the interference power at the input so that $N_{in} = 1$, formula (3) can be rewritten:

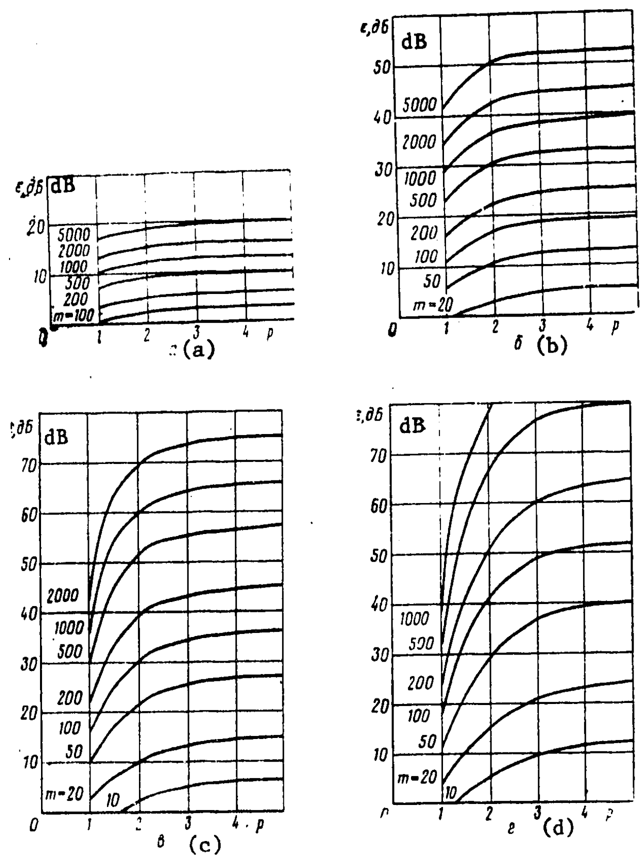
$$\varepsilon = \frac{K}{9N_{\text{вых out}}} = \frac{2\pi K}{9\int\limits_0^\infty G_N(\omega)H(\omega)\,d\omega}. \tag{5}$$

It follows from (5) that the calculation of ε can be reduced to finding the "uncompensated" interference power at the IPC output, the value of which depends on how precisely the rejection compensation filter is matched to the envelope of the interference spectrum. If the sounding signal is periodic and the interference is narrow band, the computation of ε from formula (5) can be simplified by assuming that the signal-to-noise ratio for the composite signal will be the same as for the parts of it in a frequency range of zero to F_π or F_π to $2F_\pi$, etc. The narrow band nature of the interference makes it possible to use the first term from an expansion in a Taylor series instead of a precise function for $H(\omega)$ in the integral of (5), for example, for the gaussoid:

$$N_{\text{вых}\,1} = \frac{1}{2\pi}\int\limits_0^\infty G_{N1}(\omega)H(\omega)\,d\omega = \frac{1}{2\pi}\int\limits_0^\infty \frac{4\sqrt{\pi a}}{\omega_0}\times$$

$$\times \exp\left\{-a\left(\frac{\omega}{\omega_0}\right)^2\right\}\left(\frac{\omega}{2F_\text{п}}\right)^{2p}d\omega = \frac{\pi^{2p}(2p-1)!!}{(2a)^p}\left(\frac{\omega_0}{\omega_\text{п}}\right)^{2p}, \tag{6}$$

where (2p - 1)!! is the factorial of the odd numbers; a is a dimensionless parameter determined experimentally [1].



The subnoise visibility coefficient as a function of the subtraction multiplicity factor for an exponential approximation of the power spectrum of the interference when $\nu = 2$ (a); $\nu = 3$ (b); $\nu = 4$ (c) and $\nu = 5$ (d).

When computing the residual interference power in (6), and thereafter, it is assumed that the IPC circuit incorporates a heterodyne stage for "wind velocity compensation", which makes it possible to combine the zeros of the frequency characteristic $H(\omega)$ with the "wind drift" frequencies: $(\alpha F_\pi + f_1)$, where $\alpha = 1, 2, \ldots$. By substituting (6) in (5) for a p-multiple IPC, we obtain:

$$\varepsilon_1 = \frac{K}{9\pi^{2p}}\left(\frac{\omega_n}{\omega_0}\right)^{2p}\frac{(2a)^p}{(2p-1)!!}. \tag{7}$$

An analysis of function (7) makes it possible to draw the following conclusions.

1. The results of calculations performed using formula (7) for p = 1 and p = 2 are in good agreement with the known results of [1, 2].

By using formula (5), one can compute the residual interference power for the case of an exponential approximation of its energy spectrum:

$$\begin{aligned} N_{\text{вых}\,2} &= \frac{1}{2\pi}\int_0^\infty G_{N2}(\omega)H(\omega)\,d\omega = \\ &= \frac{1}{2\pi}\int_0^\infty \frac{1{,}7\nu\sqrt{2a}\sin\frac{\pi}{\nu}}{\omega_0\left[1+\left|\frac{\omega}{\Omega}\right|^\nu\right]}\left(\frac{\omega}{2\pi}\right)^{2p}d\omega = \\ &= \frac{(\pi\omega_0)^{2p}}{(0{,}85\omega_n)^{2p}(2a)^p}\,\frac{\sin\frac{\pi}{\nu}}{\sin\frac{2p+1}{\nu}\pi}. \end{aligned} \tag{8}$$

Integral (8) exists when $\nu \geq 2p + 1 > 0$, where $\Omega = 2\pi F_e = \omega_0/0.85\sqrt{2a}$ (or $p \leq (\nu - 1)/2$). Substituting (8) in (5), we find:

$$\varepsilon_2 = \frac{K}{9\pi^{2p}}\left(\frac{0{,}85\omega_n}{\omega_0}\right)^{2p}(2a)^p\,\frac{\sin\frac{2p+1}{\nu}\pi}{\sin\frac{\pi}{\nu}}. \tag{9}$$

If the subtraction multiplicity factor is $p > (\nu - 1)/2$, integral (8) tends to infinity, something which is evidence of the impermissibility of the simplifications which were made, since the residual noise power for a passive circuit cannot exceed the power at its input. In this case, (5) can be calculated numerically.

The results of calculations of ε on a computer are shown in Figure 1 for various powers ν and subtraction multiplicity factors p. The parameter for the curves of Figure 1 is the ratio $m = F_{\pi}\sqrt{a}/f_0$.

An analysis of the curves of Figure 1 shows the following.

1. When the interference spectrum is approximated by a gaussian curve, the value of ε increases monotonically with an increase in the subtraction multiplicity factor. When the value of the parameter $m \leq 20$, the gradient of ε amounts to about 10 dB/subtraction. When $m > 20$, it increases up to 25 dB/subtraction and more.

2. When the interference spectrum is approximated by an exponential function, the tendency for the saturation of ε with an increase in the subtraction multiplicity factor is manifest all the more strongly, the smaller the exponent is or the decrease in the interference spectrum. When $\nu = 2$, the effectiveness of IPC is poor, and practically does not depend on the subtraction multiplicity factor. When $\nu = -3$ to -4, an increase in p from one to two yields a substantial gain in the value of ε, however, a further increase in the subtraction multiplicity factor does not lead to a marked increase in the subnoise visibility coefficient. When $\nu \geq 5$, increasing the subtraction multiplicity factor up to a value of $p = 3$ leads to a proportional increase in ε.

3. To improve the subnoise visibility of signals against a background of clutter from local terrain, the sea and weather formations by means of IPC, it is desirable to increase the repetition rate of the sounding signals.

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THE SYNTHESIS OF AN ARTIFICIAL APERTURE FOR MICROWAVE RADIO SIGNALS REFLECTED FROM THE SEA SURFACE

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[Paper by B.S. Mush]

[Text] Recommendations for the performance of experiments which ascertain the properties of a reflecting sea surface (MP) are given on the basis of an analysis of a formula to estimate a reflecting surface function (FOP), which takes the form of a linear superpositioning of simple waves and the returns of a radar with a synthesized antenna aperture. Results of experiments are given which show that the sea surface produces a return signal which is suitable for effective coherent accumulation by virtue of the wave nature of its reflecting surface function.

Introduction

A characteristic feature of the sea surface from the viewpoint of radar operation with a synthesized antenna aperture (SAR) is its timewise variability: the travel of an element of the sea surface in the direction of the line of viewing over a time equal to the synthesis interval is much greater than the radar wavelength λ. If this travel is treated as chaotic, then it should lead to signal errosion [1], a result of which is a loss of image quality. Because of the complexity of the phenomenon a model of synthesized aperture radar operation for the sea surface should be based on experimental data, which should be as simple and as universal as possible. Experimental studies performed with the participation of the author of this paper in the X-band* showed that the signals reflected

*The designations of the frequency bands are given in accordance with the table given in [5].

from the sea surface can be subjected to efficient coherent accumulation, as a result of which, radar images of the sea surface were obtained (figure 1) and similar results were obtained in the L-band [2, 3, 4].

1. The Theory and Principles for Obtaining Images of Wave Type Reflecting Surface by Means of Synthesized Aperture Radars

The following presentation has the purpose of ascertaining the quantitative relationship between a synthesized aperture radar signal and a timewise variable function which describes the properties of reflecting surfaces. These expressions were used to analyze signals reflected from the sea surface, and were also employed in experimental studies. The signal s(t), reflected from the surface, at the input to the radar processing system can be represented in the form:

$$s(t)=\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}K(X(t)-x)Q\left(t+\frac{2}{c}(Y(t)-y)\right)\exp(-2iky)\times \times f(x,y,t)\,dx\,dy+n(t), \quad (1)$$

where

$$K(z)=G^{2}\left(\frac{z}{Y(0)}\right)\exp-ik\frac{z^{2}}{Y(0)};$$

and the integration is carried out in a rectangular system of coordinates (x, y), in which case, the direction of the main lobe of the radar directional pattern G at the point in time p = 0 is collinear with the y axis; X(t) and Y(t) are the projections of the radius vector $\vec{R}(t)$ on the x and y axes, where the end of the radius vector coincides with the phase center of the radar antenna (Figure 2); Q(t) is a complex function which describes the sounding signal; f(x, y, t) is the complex reflecting surface function (FOP); k is the wave number ($k = 2\pi/\lambda$); n(t) is the additive noise of the system. If the FOP is not a function of time, equation (1) can be treated as an integral equation of the first kind with a "near difference" nucleus [6] and an unknown function f, the estimate of which is determined by means of solving integral equation (1). If the FOP depends substantially on time, then (1) loses the solution uniqueness, with the exception of one special FOP case, specifically, a wave type function [4]:

$$f(x,y,t)=f(x-x_{\tau},\,y-y_{\tau}),$$

where x_T and y_T are the coordinates of the vector $\vec{\rho}_{\tau}=\int_{t_0}\vec{w}_{M}(t)\,dt;\ \vec{w}_{M}(t)$; $\vec{w}_M(t)$ is the variable group velocity vector of the wave. The determination of such FOP's as waves is based on the fact that this concept does not requite the obligatory travel at a velocity $\vec{w}_M$ of the material carriers of the waves, but only their form, and uses the concept of an ideal wave as that of the travel of "perturbations in space at a constant velocity without a change in their form" [7]. We shall go ahaed with a further quantitative analysis for a simplified case, in which the sounding signals are simulated by a train of δ-pulses, where δ is a Dirac function:

$$Q(t)=\sum_{n=-N/2}^{N/2}\delta(t-nT_{п}),$$

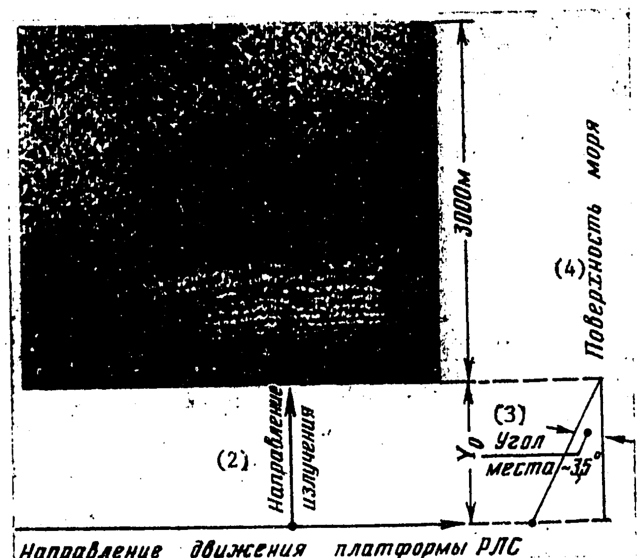

Figure 1. The radar image of a sea surface obtained by means of artificial aperture synthesis.

Key: 1. Direction of travel of the radar platform;
2. Direction of radar transmission;
3. Elevaticn angle ≈3.5°;
4. Sea surface.

where T_π is the repetition period of the sounding signals; N is the number of pulses used in the synthesis. Then equation (1) is simplified, specifically:

$$\int_{-\infty}^{\infty} K(X(t)-x) f(x, \bar{y}, t)\, dx = s(t) \exp 2ikY(t) + n(t), \tag{2}$$
$$\bar{y} = \frac{c}{2}\theta + Y(t), \quad \theta = t - nT_n.$$

Let f be independent of t, and then the solution of (2) for a fixed value of y = $\bar{y}$ = const. is determined by an integral operator with a nucleus $\mathscr{R}$ [4]:

$$\hat{f}(x, \bar{y}) = \int_{(T)} \mathscr{R}(x - X(t), \bar{y}) s(t) \times \exp 2ikY(t)\, dt, \tag{3}$$

where T is the synthesis interval. We shall call the function of the form $\hat{f}(x, \bar{y})$ the FOP estimate. The nucleus of the transform (3) can form a statistically optimal one in the sense of the mean square deviation of the Wiener-Kolmogorov operator [8]:

$$\mathscr{R}(\xi) = \int_{-\infty}^{\infty} \frac{\tilde{K}^*(p) F(p)}{N(p) + F(p) |\tilde{K}(p)|^2} \exp i\xi p\, dp, \tag{3.1}$$

where F(p) and N(p) is the energy spectrum of the functions f and n; the symbol ($\tilde{\cdot}$) designates the Fourier transform of the corresponding function, while the symbol (*) designates the complex conjugate. Operator (3.1) yields the dispersion of the estimate of the FOP, σ_f, proportional to the integral:

$$\sigma_f \sim \int_{-\infty}^{\infty} \frac{N(p) F(p)}{N(p) + F(p) |\tilde{K}(p)|^2} dp. \tag{3.2}$$

In the case where the following equality has a physical meaning,

$$N/F \gg |\tilde{K}(p)|^2$$

throughout the entire frequency range, operator (3.1) forms the widely disseminated "matched filter". In the case of rectilinear and uniform motion of the radar platform:

$$X(t) = w_{nx} t = w_n t \cos\beta,$$
$$Y(t) = w_{ny} t = w_n t \sin\beta, \tag{3.3}$$

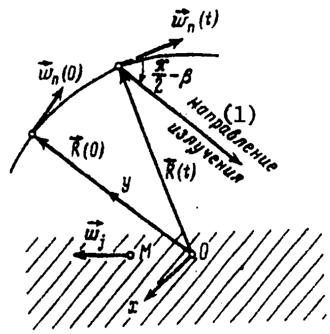

Figure 2. The geometric relationships of the system (x and y are the coordinate axes; w_π is the velocity vector of the radar platform; w_j is the velocity vector of the j-th wave of the FOP [surface reflection function]; $\vec{R}(t)$ is the radius vector of the phase center of the antenna).

Key: 1. Direction of radar transmission.

where $w_{\pi x}$ and $w_{\pi y}$ are the projections of the $\vec{w}_\pi$ vector on the x and y axes, while β is the angle between the y axis and the normal to the vector $\vec{w}_\pi$, estimate (3.3) has the following form:

$$\hat{f}(x,\bar{y}) = \int_{-\infty}^{\infty} K^*(p)\,\tilde{s}(p-2k\,\mathrm{tg}\,\beta)\exp ipx\,dp,$$

$$\tilde{s}(p) = K(p+2k\,\mathrm{tg}\,\beta)\,\tilde{f}(p+2k\,\mathrm{tg}\,\beta,\ \bar{y}),$$

where $\tilde{f}(p, \bar{y})$ is a one-dimensional Fourier transform of the function $f(x, \bar{y})$. We shall also introduce the following definition of the estimate uncompensated with respect to the angle β:

$$\hat{f}_{\beta,\beta_0}(x,\bar{y}) = \int_{-\infty}^{\infty} K^*(p)\,\tilde{s}(p-2k\,\mathrm{tg}\,\beta_0)\exp ipx\,dp,$$

obtained by means of applying the operator R, corresponding to the angle β_0, to the signal obtained for a transmission direction equal to β. It follows from what has been said above that:

$$\hat{f}_{\beta,\beta_0}(x,\bar{y}) = \int_{-\infty}^{\infty} K^*(p)\,K(p+2k(\mathrm{tg}\,\beta-\mathrm{tg}\,\beta_0))\times$$
$$\times\,\tilde{f}(p+2k(\mathrm{tg}\,\beta-\mathrm{tg}\,\beta_0),\bar{y})\exp ipx\,dp.$$

Let f(x, y, t) be a function of time and one which can be represented as the linear superposition of M wave FOP's, f_j, with a wavelength Λ_j and a group propagation velocity of $\vec{w}_j = \vec{w}(\Lambda_j)$:

$$f(x,y,t) = \sum_{j=0}^{M-1} f_j(x-w_{jx}t,\ y-w_{jy}t),$$

where w_{jx} and w_{jy} are the projections of the vector $\vec{w}_j$ onto the x and y axes, and then the signal s(t) has the form

$$s(t)=\sum_{j=0}^{M-1}\int_{-\infty}^{\infty}K(w_{nx}t-x)f_j(x-w_{jx}t,\bar{y}-w_{jy}t)\exp-2ikw_{ny}t\,dx$$

or

$$s(t)=\sum_{j=0}^{M-1}s_j(t),$$

$$s_j(t)=\int_{-\infty}^{\infty}K(\chi_{jt}-\xi_j)f_j(\xi_j,\bar{\eta}_j)\exp-2ikH_{jt}\,d\xi_j,$$

where

$$\xi_j=x-w_{jx}t;\quad \chi_{jt}=(w_{nx}-w_{jx})t;$$

$$\eta_j=y-w_{jy}t;\quad H_{jt}=(w_{ny}-w_{jy})t.$$

Thus, when M = 1, the synthesis procedure is similar to that which is done for the case of timewise constant FOP and a known value of the sum $\vec{w}_\pi+\vec{w}_j$. In the general case:

$$\tilde{s}(p)=\sum_{j=0}^{M-1}\frac{1}{w_{nx}-w_{jx}}\tilde{K}\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}}\right)\times$$

$$\times\tilde{f}_j\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}},\bar{\eta}_j\right),$$

$$p_j=\frac{w_{nx}}{w_{nx}-w_{jx}}p,$$

and the FOP estimate which is uncompensated with respect to β when $\beta_0=0$ is:

(4)
$$\hat{f}_{\beta,0}(x,\bar{y})=\sum_{j=0}^{M-1}\int_{-\infty}^{\infty}\tilde{K}^*(p_j)\tilde{K}\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}}\right)\tilde{f}_j\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}},\ \bar{\eta}_j\right)\exp i\frac{w_{nx}-w_{jx}}{w_{nx}}p_jx\,dp_j.$$

We shall consider the integral brightness function of the FOP estimate uncompensated with respect to β, or simply the brightness function, defined as:

(5) $$J(\beta)=|\overline{f_{\beta,\beta_0}(x,y)}|^2,$$

where the symbol $(\bar{\cdot})$ indicates statistical averaging with respect to the ensemble f_{β,β_0}. The use of (4) and (5) following the transformations yields:

$$J(\beta)=\frac{1}{w_{nx}}\int_{-\infty}^{\infty}|K(p)|^2 B_s(p)\,dp,$$

where $\tilde{B}_s$ is the energy spectrum of s. Considering the fact that $|\tilde{K}(p)|$ is satisfactorily bounded by the limits $[-p_2, p_2]$, where $p_2 > 0$, we have:

(5.1) $$J(\beta)\propto\frac{1}{w_{nx}}\int_{-p_2}^{p_2}B_s(p)\,dp.$$

The following formal relationship exists between the energy spectrum $\tilde{B}_\psi(p)$ of the random function $\psi(t)$ and the function $|\tilde{\psi}(p)|^2$:

(6) $$|\overline{\Psi(p)}|^2=\delta(0)B_\psi(p),$$

while at the same time, the function $\overline{|\tilde{s}(p)|^2}$ can be represented in the form:

(7) $$\overline{|\tilde{s}(p)|^2}=\sum_{j=0}^{M-1}\frac{1}{(w_{nx}-w_{jx})^2}\left|K\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}}\right)\right|^2\times$$
$$\times\overline{\left|f_j\left(p_j-2k\frac{w_{ny}-w_{jy}}{w_{nx}-w_{jx}}\right)\right|^2}$$

assuming the orthogonality of f_j as independent random functions:

$$\overline{f_j f_k^*}=\begin{cases}0, & \text{если (if) } j\neq k,\\ |f_j|^2, & \text{если } j=k.\end{cases}$$

Taking (5.1), (6) and (7) into account, we have:

$$J(\beta)=\sum_{j=0}^{M-1}\frac{w_{\Pi x}}{(w_{\Pi x}-w_{jx})^3}\int\limits_{-p_2\frac{w_{\Pi x}-w_{jx}}{w_{\Pi x}}}^{p_2\frac{w_{\Pi x}-w_{jx}}{w_{\Pi x}}}\left|\bar{K}\left(p_j-2k\frac{w_{\Pi y}-w_{jy}}{w_{\Pi x}-w_{jx}}\right)\right|^2\times$$
$$\times F^{(j)}\left(p_j-2k\frac{w_{\Pi y}-w_{jy}}{w_{\Pi x}-w_{jx}}\right)dp_j,$$

where $F^{(j)}(p_j)$ is the energy spectrum of the j-th component of the FOP. Given the condition that $w_{\pi x} \gg w_{jx}$:

(8)
$$J(\beta)\simeq\frac{1}{w_{\Pi x}^2}\sum_{j=0}^{M-1}\int\limits_{-p_2-2k\left(\mathrm{tg}\,\beta-\frac{w_{jy}}{w_{\Pi x}}\right)}^{p_2-2k\left(\mathrm{tg}\,\beta-\frac{w_{jy}}{w_{\Pi x}}\right)}F^{(j)}(p_j)\,dp_j.$$

An analysis of expression (8) shows that in the case of a single component, either moving or stationary FOP (M = 1), the width of the brightness function $J(\beta)$ is determined by the width of the antenna directional pattern and the width of the FOP energy spectrum expressed in angular units: the relationship between the spatial frequency and the angular units is defined by the equality:

$$p\lambda = 4\pi\beta.$$

The brightness function of a moving single component surface has a maximum which is shifted relative to the maximum of the brightness function of a stationary surface by the amount $w_{jy}/w_{\pi x}$. If during the measurement of this shift, under conditions where there is a lack of apriori knowledge of the direction $\vec{w}_j$, this shift is not detected, then this means that the vectors $\vec{w}_j$ and $\vec{w}_\pi$ are colinear and it is necessary to measure the brightness function of the moving surface at a different value of β_0. If the width of $F^{(j)}$ is much less than the quantity $2p_2$, then the width of the brightness function $\Delta\beta$ is equal to:

(8.1)
$$\Delta\beta=\beta_a+\frac{1}{w_{\Pi x}}(w(\Lambda_{\text{макс}})-w(\Lambda_{\text{мин}}))+\frac{\lambda}{2}\left(\frac{1}{\Lambda_{\text{мин}}}-\frac{1}{\Lambda_{\text{макс}}}\right),$$

where Λ_{min} and Λ_{max} are the minimum and maximum wavelengths in the set of M wave FOP's, given that condition that the function $w(\Lambda_j)$ is nondecreasing, from which it follows that the presence of dispersion properties in a complex FOP, which can be represented by the superposition of simple waves, and the amount of difference between the ultimate wave numbers of the FOP waves have an impact on the widening of the brightness function. It follows from this same formula that the width of the brightness function of a stationary surface for synthesis intervals close to the first Fresnel zone, is basically governed by the width of the actual antenna directional pattern β_a. That FOP for which there exist components f_j with any finite group velocity $\vec{w}_j$, can be considered to be a completely noise-like FOP.

2. The Application of the Brightness Function to the Study of the Background Properties of the Sea Surface

We shall employ the hypothesis that the sea FOP can be represented by a set of wave FOP's and that the function $w(\Lambda_j)$ obeys a hydrodynamic dispersion law similar to that which exists for waves at a water surface [9]:

$$w(\Lambda_j) = \left(\frac{g}{2\pi}\Lambda_j + \frac{2\pi\alpha}{\rho}\frac{1}{\Lambda_j}\right)^{1/2}, \tag{9}$$

where g is the free fall acceleration; α is the surface tension cieffficient; while ρ is the density of the sea water. The first part of the hypothesis has a rather general form and is hardly in need of a serious check, since any physically feasible space-time function can be represented as a set of simple waves, for example, by means of expansion in a two dimensional Fourier series. The second part of the hypothesis though is in need of a check. We shall introduce the symbol:

$$\Lambda_{\text{мин}} = \Lambda_{\text{макс}} - \delta\Lambda, \qquad \Lambda_{min} = \Lambda_{max} - \delta\Lambda,$$

with the condition that $\delta\Lambda/\Lambda_{max} \ll 1$, we have the following taking (8.1) into account:

$$\Delta\beta = \beta_a + \left(\frac{\dfrac{g}{2\pi} + \dfrac{2\pi\alpha}{\rho\Lambda^2_{\text{макс}}}}{2w_{\text{вх}}w(\Lambda_{\text{макс}})} + \frac{\lambda}{2\Lambda^2_{\text{макс}}}\right)\delta\Lambda. \tag{10}$$

In formula (10), the second term corresponds to an increase in the width of the brightness function of the "sea" as compared to "dry land". The representation of the FOP of the sea with a finite set of wave components is taken as the basis for the treatment of the experimental data presented below.

3. Experimental Studies of the Specific Features of the Brightness Function of Radars with Synthesized Apertures for Signals Reflected from the Sea Surface

The experimental studies consisted in recording the coherent signals reflected from a sea surface and a dry land surface, with their subsequent digital processing. The radio signals were generated by means of a moving directional radio emission source in the centimeter band with horizontal polarization for a directional pattern width of about 3°. The elevation angle amounted to 2.5 to 10°: specifically, the image of the sea in Figure 1 was obtained for an elevation angle of about 3.5° and is an estimate of a FOP of the form of (3) in a synthesis interval equal to the width of the first Fresnel zone. The processing of the recorded signals consisted in computing the FOP estimate uncompensated with respect to β of the form of (4) in a range equal to the first Fresnel zone, and the subsequent calculation of the brightness function (5). In order for it to be possible to compute the brightness function, the signals were recorded with a slow change in the angle β. The precision in the determination in the angle β was about 6'. The width of the brightness function was defined at the ≈ 20 dB level.

The experiment which was performed ascertained the following facts.

1. The width of the brightness function for sea and dry land differs insignificantly. Thus, in one of the series of experiments in which the state of the sea was about 4 points [4 points on a 12 point scale], the following results were obtained:

For the sea, $\Delta\beta = 4.560 \cdot 10^{-2}$ radians
For the dry land $\Delta\beta = 4.656 \cdot 10^{-2}$ radians

for an antenna pattern width of $\approx 4.260 \cdot 10^{-2}$ radians. The time for one measurement of the brightness function amounted to 60 synthesis periods. Then the individual realizations of $J(\beta)$ were averaged for 20 values. If formula (10) is used, then we obtain $\delta\Lambda = 0.92$ m for Λ_{max}.

2. The measurement of the position of the maxima of the function $J(\beta)$ for sea and dry land shows that there exists a shift in them Thus, in one of the experiments for sea wave agitation of about 3 points [3 points on a 12 point scale], a shift by an amount of about $2.4 \cdot 10^{-2}$ radians was observed, a figure which amounted to more than half of the width of β_a and corresponds to a radial velocity of the FOP of about 2.9 m/sec. The value obtained is rather close to the value of the velocity of the

corresponding sea gravitational waves computed from formula (9). In the case of special observations which corresponded to the collinearity of the vectors $\vec{w}_\pi$ and $\vec{w}_1$, the shift was not perceptible. However, for a change in the angle of the initial setting of the antenna β_0 by about 20°, a marked shift in the maxima of $J(\beta)$ for sea and dry land appeared.

3. It was noted that the size of the maximum of the brightness function increased with an increase in the degree of sea agitation, however, this phenomenon occurred only when the wind strengthened and is poorly correlated with the presence of wave crests. Stable synthesized images appeared with sea wave agitation of more than 2 points and wind higher than 2 m/sec.

Conclusions

1. The sea surface in the centimeter band and with an observation time of less than one second for the case of artificial aperture synthesis produces a radar picture which is premoninantly of a single wave structure, moving at a constant velocity, and in this sense is a result of effective coherent accumulation.

2. The phenomenon of shift with respect to the azimuthal angle of the brightness function of the synthesized images of the sea relative to that for dry land can be explained by the motion of gravitational sea waves.

3. Based on what has been presented above, it can be presupposed that a specific feature of synthesized images of the sea is the presence of brightness contrasts of the individual surface sections with an anomalous group velocity of motion, something which opens up new possibilities in mapping the surface of the sea.

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